rf design

engineering principles and practices

December 1989

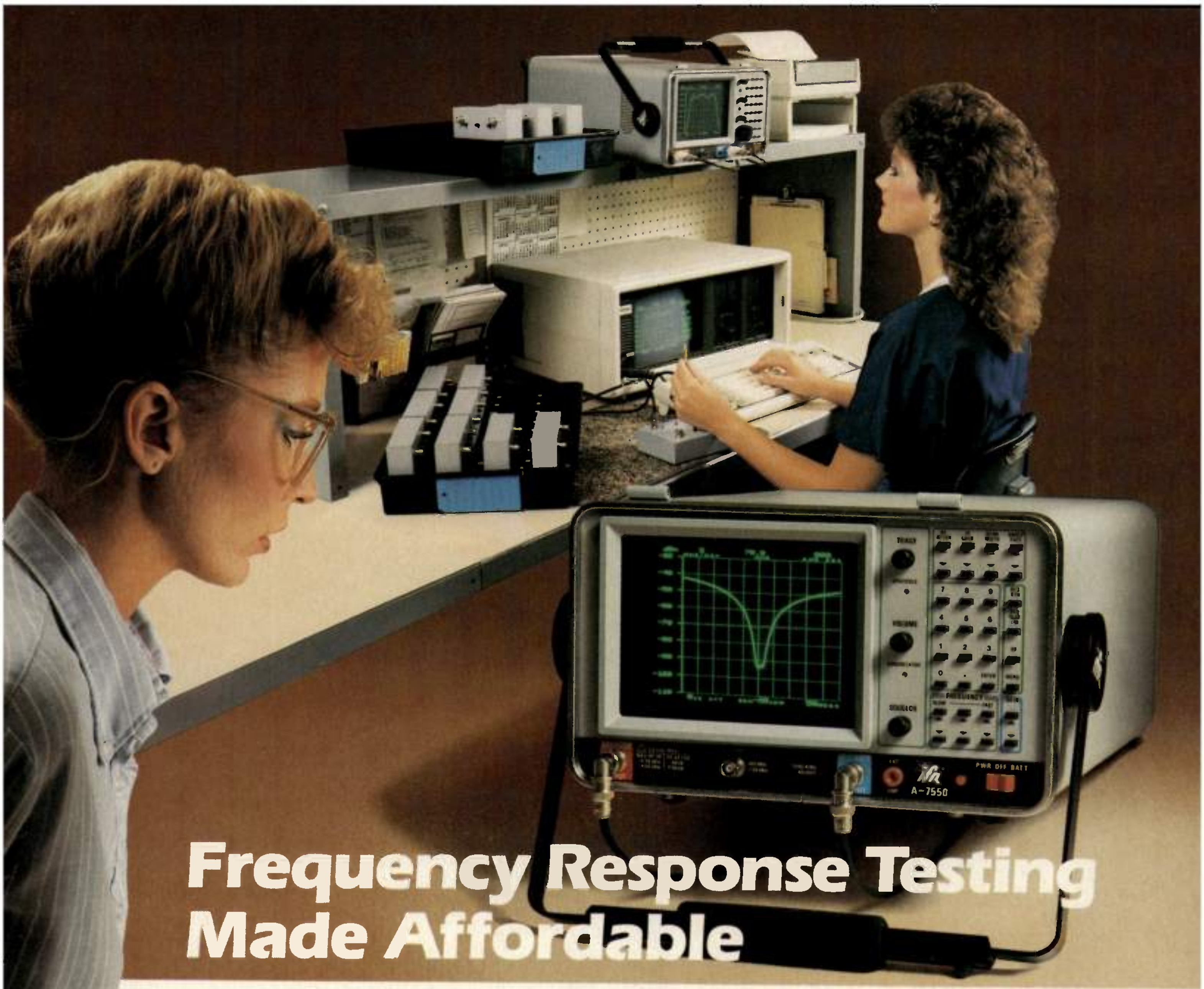
Complex Signals
— A Tutorial Series

Industry Insight

RF Cables and Connectors

Plus —

1988-1989 Article Index



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INFO/CARD 2

TWEAKING

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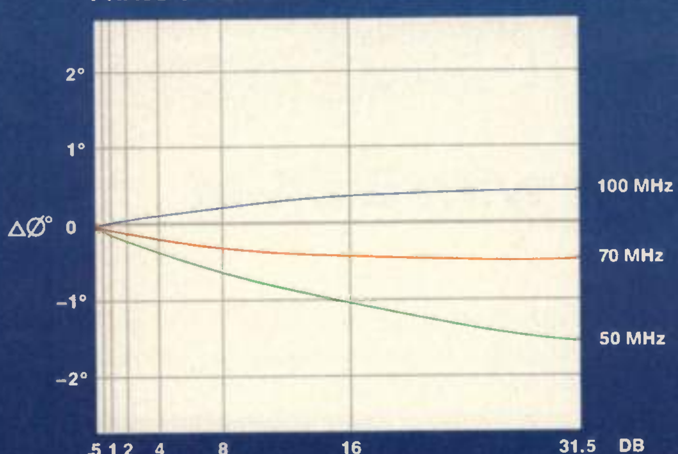
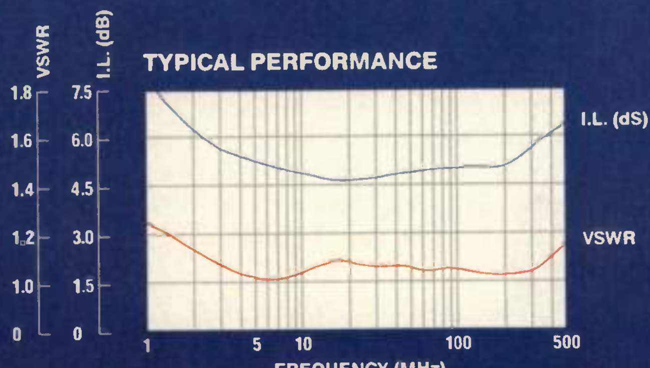
PHASE CHANGE VS. ATTENUATION



TYPICAL PERFORMANCE



OPERATING CHARACTERISTICS

PARAMETER	MIN	TYP	MAX	UNITS	CONDITION
SWITCHING TRANSIENTS		54.0	100.0	mV	PEAK VALUE
TRANSITION TIME		5.0		nSEC	90%/10% or 10%/90% RF
SWITCHING SPEED		20.0	35.0	nSEC	50% TTL to 90%/10% RF
MONOTONICITY GUARANTEED					
INSERTION LOSS		4.9	6.0	DB	
△ PHASE		0.4	±1.0	DEG	70 MHz
ATTENUATION: RANGE	0		31.5	DB	
STEPS					0.5, 1, 2, 4, 8, 16dB
VSWR		1.08	1.35/1		

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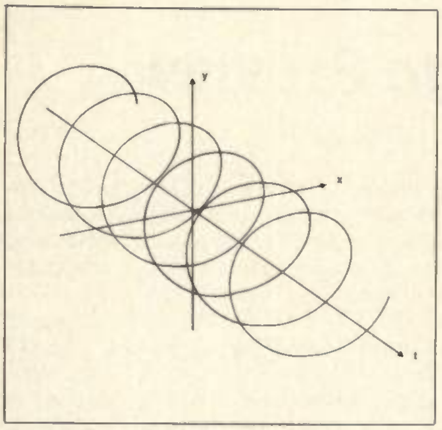
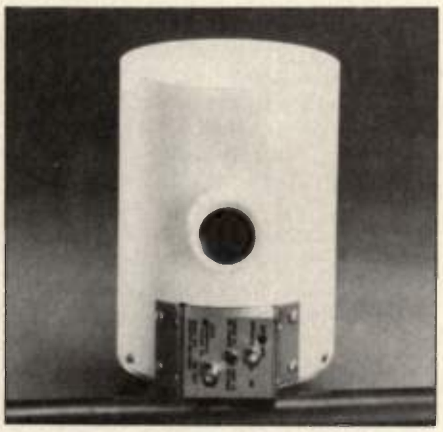
rf design

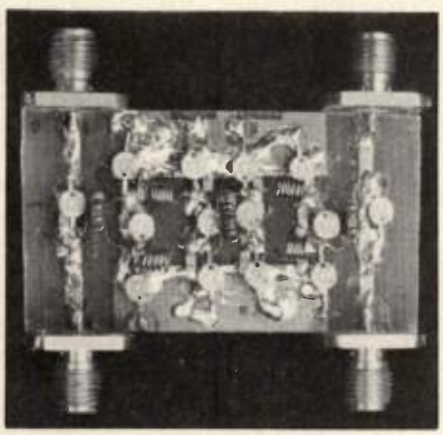
December 1989



Page 27 — Complex Signals



Page 47 — Bowtop Antenna



Page 51 — Quadrature Coupler

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rf editorial

A 1990 *RF Design* Preview

By Gary A. Breed
Editor

It's the time of year to look ahead to 1990 and see what we have planned for you and the 40,000 other RF engineers who receive *RF Design*. As you read about our plans, think about contributing an article, submitting a contest entry, or offering your ideas on the topics we'll be featuring.

First, an exciting new addition is a question-and-answer column for solving design problems. You are invited to ask your most frustrating questions (we won't publish names in the magazine, so don't be embarrassed to ask). Members of our editorial review board, and other experts in the RF industry will provide the answers. The first installment of this new feature is in January.

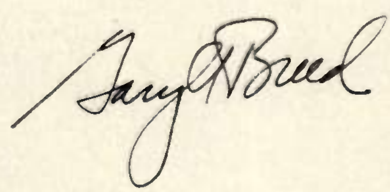
In 1989, we started our Industry Insight column, examining technology and marketplace trends. We've had an excellent response to these reports, and learned a lot ourselves. 1990 will see new areas explored, including oscillators, low-cost instruments, budget software, low noise transistors, MMICs, EMI control products, and more.

Of course, the key engineering topic each month is highlighted in our Featured Technology section. The first six months of 1990 look like this: data communications in January; then a three-part series on communications systems, with transmitters covered in February, receivers in March, and antenna systems in April. May will feature frequency synthesis, and June will bring us to mid-year with a look at computer synthesis and design techniques.

The Fifth Annual RF Design Awards Contest is now taking entries, and will have results published in July. With more prizes than ever the contest is sure to attract another crop of fascinating design solutions. We'll continue what we started last July, publishing an entry from the previous contest in every issue, on handy perforated cards to remove and put in your notebook.

As *RF Design* continues to grow (1989 was our best year ever), we are increasing the number of subscribers, and we are adding more features to help you stay on top of the engineering field. The past couple of years have been a time of some uncertainty in the RF market, especially with reduced government spending. But, there are also lots of new opportunities in military, commercial, and consumer markets that are cause for optimism. We have good reason to take an optimistic viewpoint, because we feel that our success reflects the vitality of the RF industry.

Keep in touch, this is going to be another great year for RF!



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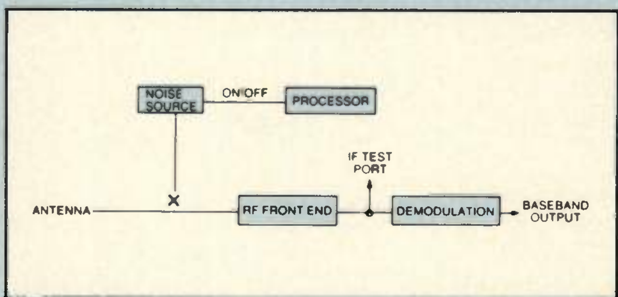

Noise sources can also be used in setting minimum detectable signal level, establishing dynamic range, or setting a reference for go/no-go testing. Frequency response can be measured at a test port prior to the demodulation circuitry. Since noise sources provide constant energy across the receiver bandwidth, frequency response can be measured instantaneously — without tuning an input signal.

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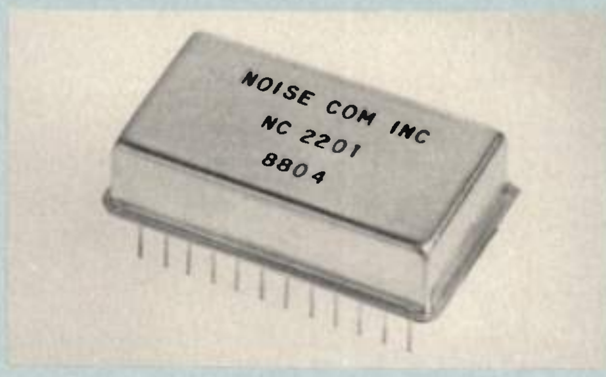

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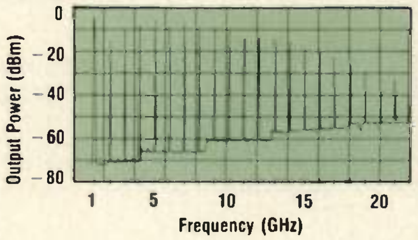
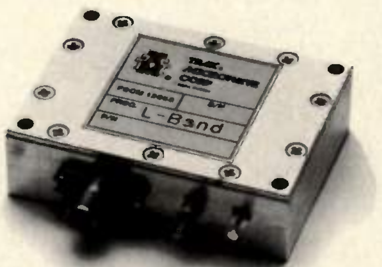
Typical Power vs. Frequency Curve





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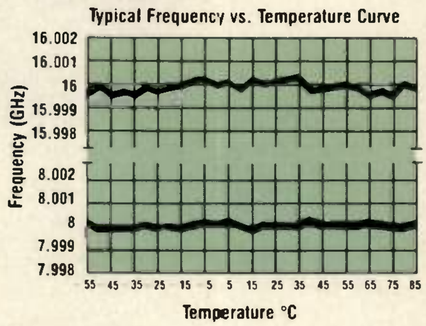
Our 6720 Series Ultra-stable DSOs deliver + 16dB minimum, and are stable to ± 500 KHz at frequencies from 8-18GHz.

These and other types of oscillators, such as XCOs, are used in many systems such as RWRs, DF systems, surveillance, Elint, simulators, decoys, and others. For high performance in small envelopes, EW systems throughout the Western forces use TRAK oscillators.

Ultra Stable D.S.O.



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Circulators & Isolators

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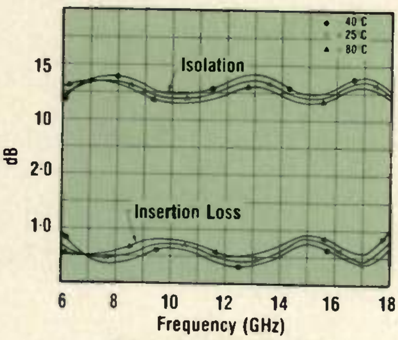
Typical Performance INFO/CARD 11



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INFO/CARD 12

YIGs

INFO/CARD 13

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INFO/CARD 14

Call or write for our Free Components Catalogs. See EEM or MPDD for other TRAK military products.

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rf letters

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

Filter Design Discussion

Editor:

I read William Lurie's comments *(RF Design*, October 1989) on my article about Chebyshev filters *(RF Design*, August 1989) with much interest. It's reassuring to discover that someone not only read the article, but took the trouble to comment on it. However, at least one of his criticisms appears to be based on a misconception.

Most filter designers would agree that no single type is always the best choice, but Mr. Lurie comes perilously close to asserting that an elliptic filter, or some similar type of filter with finite zeros in the stopband, is just such a panacea. In fact, a Chebyshev filter is often the best choice when the requirements are relatively undemanding. For example, a filter with maximum passband ripple of 0.5 dB and a 0.5 to 60 dB shape factor of 2.0:1 could theoretically be a five-pole elliptic or a seven-pole Chebyshev filter. Either would require seven components, but the Chebyshev filter would have the advantage of monotonically increasing stopband attenuation.

The statement that "bandpass filter designs are always based on doubly terminated, minimum insertion loss prototypes" was intended to apply only to Chebyshev and Butterworth L-C filters, the subject discussed in the article. In addition to the examples given by Mr. Lurie, it obviously isn't true for active and digital. Even so, it should have been qualified ("almost always").

A more serious criticism, charitably overlooked by Mr. Lurie, is the questionable use of the term "insertion loss." This usually refers to passband attenuation caused by the finite Q of the filter components, not to loss caused by an impedance mismatch between the source and load. In retrospect, "insertion loss" should have been called "mismatch loss," or something similar.

Jack Porter
Cubic Defense Systems
San Diego, California

RF Expo East 89: Diversity and Depth

This year's RF Expo East, held in Atlantic City, offered RF engineers a chance to view the latest products and technologies, in a setting affording the opportunity for exchange of ideas between fellow engineers. Exhibitors were pleased to find a serious crowd of engineers with the goal of finding products to solve specific design problems. And attendees were presented with a technical program offering both fundamental design information and innovative design techniques.

With few exceptions, the technical program was praised by attending engineers for its excellence and relevance to their work. Technical Program Chairman Andy Przedpelski commented, "The papers seemed to be directly applicable to everyday problems. However, there were also some new, not yet widely used techniques presented." The largest crowds attended sessions on receiver design, frequency synthesis and phase-locked loops.

Although all the presentations were well-received, a number of papers got particularly good marks from the audience. Three of these were 1989 RF Design Awards Contest entries: "A Reference-Cancelling Phase/Frequency Detector" by Dan Baker; "An RF Active Elliptic Filter" by Eric Kushnick; and "An Unconventional Varactor-Tuned Filter" by Gary Thomas.

Other papers of note included the "Antenna Principles" tutorial by Benjamin Rulf; "Microcomputer-Compensated Crystal Oscillator With Self-Temperature-Sensing," presented by John Vig; and "Design, Implementation, and Test of a Wideband HF

Receiving Subsystem" by John Link and Harry Lenzing. These were among the best of many excellent papers.

The three special courses, Fundamentals of RF Circuit Design: Parts I and II and Computer-Aided Filter Design, were again well-received and well-attended. Instructors Les Besser (complete with Bay area earthquake debris) and Randy Rhea once again got high marks for their presentations.

Engineers attending RF Expo East 89 were given a hands-on look at the latest products being offered by more than 80 exhibiting companies. Among the highlights of the show was the announcement of Wavetek's Model 2410R ruggedized signal generator, covering 10 kHz to 1100 MHz, with IEEE interface, −127 to +13 dBm output level, and extensive internal diagnostic capability. The unit on display at Wavetek's booth had been shipped around the world, and dragged behind a jeep for a mile at 35 mph, and still met published specifications.

Direct Conversion Technique Inc. introduced a line of portable network analyzers. DCT SNA-30 covers 1.5 to 60 MHz, SNA-300 covers 10 to 520 MHz, and SNA-900 covers 700 to 1200 MHz. These instruments measure impedance, VSWR and insertion gain/loss of active and passive RF circuits.

On display at XL Microwave's booth was the new Model 3201 source-locking microwave counter for 10 MHz to 20 GHz, with 0.1 Hz resolution. The unit has −20 dBm sensitivity and includes GPIB interface. Litton Electron Devices unveiled the D-6872 power GaAsFET, featuring +24 dBm output and 6 dB gain at 18 GHz. The D-3860 power GaAsFET, with +31.5 dBm output power and 11.5 dB linear gain at 8 GHz, was also on display.

Ehrhorn Technological Operations (ETO) premiered a microprocessor-controlled 20 kW linear power amplifier, Model 52S-21, designed for MRI applications in the 10 to 90 MHz range. Announced by Harris Semiconductor was the HA-2546 two-quadrant multiplier IC, with 30 MHz signal bandwidth, for applications in sonar, video, radar and AGC applications.

Silicon PIN diode chips, with as low as 0.05 pF capacitance and medium power handling capability, were introduced by Unitrode. Series resistance of these devices is nominally 1.5 ohms (at 100 mA), and carrier lifetime is 400 ns (at 10 mA).

RF Expo East 89 presented a strong technical program, along with the latest in product and service offerings, and has engineers looking forward to next year's show. RF Expo East 90 is set for November 13-15, 1990 at the Marriott Orlando World Center in Orlando, Fla.

DARPA Awards Six More HDTV Contracts—Six companies have been chosen by the Defense Advanced Research Projects Agency (DARPA) to develop display processor technology for high-definition television (HDTV). The team of the David Sarnoff Research Center, Sun Microsystems and Texas Instruments will work to develop a high-definition image workstation. Adams-Russell Electronics and the Massachusetts Institute of Technology (MIT) will work jointly to develop advanced compression technology, using research previously conducted at the MIT Media Laboratory. Qualcomm Inc., of San Diego, will be working on an alternative compression technique based on state-of-the-art digital communications technology. The contracts have been awarded as part of a three-year, $30 million DARPA program for funding of HDTV research. Approximately $15 million is earmarked for work on display processor technology.

Tektronix to Acquire Colorado Data Systems—At a November 14th press conference, Tektronix vice-president Richard Knight announced that his company will purchase Colorado Data Systems (CDS) of Englewood, Colo. CDS manufactures automated test systems and modular test instruments, which Tektronix feels will complement its existing product line. Details of the agreement, including any relocation plans and new company name, have not yet been announced.

Triple Function Crystal Developed—A single piezoelectric crystal has been developed which can generate both longitudinal and shear waves simultaneously as well as independently. Researchers at Queen's University and the National Research Council of Canada had theorized that such dual mode vibration was possible, but only in materials with previously unavailable properties. Working with Valpey-Fisher Corp.'s piezoelectric specialists, the Canadian researchers have developed a transducer made from lithium niobate with characteristics matching a computerized model. This transducer generates longitudinal waves, shear waves, or both longitudinal and shear waves simultaneously, at operating frequencies from 5 to 85 MHz.

The development of the new "triple function" crystal will be welcome in the field of material evaluation, where complete characterization often requires the use of both shear and longitudinal waves. With this crystal, a single ultrasonic transducer can perform the work of two or more, reducing the time required for characterization, and eliminating suspected crystal-to-crystal variability.

Repeater Stations Aided in Valdez Cleanup—IWL Communications Inc., a Houston-based firm specializing in radio and telecommunications problems, is one of the many companies that cooperated in the search for innovative ways to speed cleanup efforts in the aftermath of the Valdez, Alaska, oil spill. IWL's Rapidly Deployable Radio Repeater Stations were quickly customized to extend and enhance EXXON's offshore and inland communications during the emergency.

The first nine units of the repeater station network were built, shipped, and operating in less than a week. The solar-powered design weighs less than 100 lbs. and is totally self-contained. Power and radio systems are in weatherproof enclosures connected by a heavy-duty umbilical cord and mounted on an aluminum frame with four eye hooks for easy lift and transport. Solar panels and antenna are self-adjustable for easy alignment. Units can be set up and operating in less than 20 minutes.

Installed at East Kodiak Island, the unit pictured here is typical of the repeater network. Each station consists of three repeaters, each programmed for single-channel operation, and broadcasting a low-power synthesized RF signal through an omnidirectional fiberglass antenna. Built-in cross-band repeaters link the EXXON Kodiak Command Center (via Pillar Mountain UHF repeater) to cleanup vessels equipped with VHF marine radios and operating in the Shelikof Strait and other Kodiak waters. IWL units also extend EXXON communication coverage to U.S. Coast Guard vessels and shore stations and to remote cleanup crews equipped with hand-held portable radios.

Auto-ID System Tracks Freight and Transport Vehicles—An automatic electronic identification system developed by AMSKAN Ltd., of Melbourne, Australia, tracks the transportation of goods using radiated RF energy. The BARTAG system, a sophisticated application of the scanners used in supermarkets, relies on infra-red radiation and FM radio signals. Small electronic labels, mounted on the side of a vehicle, railcar or shipping container, emit encoded signals that are scanned some distance away by a BARTAG reader. Reportedly, the system can operate at distances of up to 32 yards and vehicle speeds up to 62 miles per hour.

Kyocera and AVX Corp in Merger—Kyocera Corp. of Japan and AVX Corp., a leading U.S. manufacturer of ceramic capacitors, have announced the signing of a definitive merger agreement, in a transaction valued at approximately $530 million. According to the agreement, all outstanding shares of AVX common stock will be exchanged for Kyocera American Depository Shares (ADS), each of which represents two shares of Kyocera common stock. This is believed to be the first merger between a Japanese and a U.S. company using exchange of stock. The transaction is subject to various Japanese and U.S. regulatory approvals, and to approval by AVX shareholders, with a meeting expected in late December of this year.

Compact Software Offers Free CAD Seminar—Free technical seminars on computer-aided design are being offered by Compact Software at a number of locations throughout the United States and Canada. The one-day technical seminars offer an opportunity to learn about the latest advances in Compact Software's CAD/CAE products. These seminars are intended particularly for managers, engineers and technicians with an interest in learning about new RF and microwave simulation tools. See the RF Calendar column on page 20 for dates and locations of upcoming seminars. Further information can be obtained by contacting: Sam Benzacar, Training Manager, Compact Software, Inc., 483 McLean Boulevard, Corner 18th Avenue, Paterson, NJ 07504. Tel: (201) 881-1200

Growth Forecast for TEMPEST Market in U.S.—U.S. expenditures for military TEMPEST equipment will hit a low point of $648 million in 1989 (down from $766 million in 1988) before climbing steadily toward $1 billion by 1994. That is the assessment of a recent report by Frost and Sullivan, Inc. According to the study, expanding use of computers in the workplace will result in increased demand for TEMPEST versions of PCs, workstations, peripherals and local area networks, while making measures like shielding entire rooms less practical. Essentially all of the projected growth will be due to what the Frost and Sullivan study refers to as a "relaxed" TEMPEST standard, in which commercial products may be minimally modified to provide some protection, but not meet full TEMPEST signal suppression requirements.

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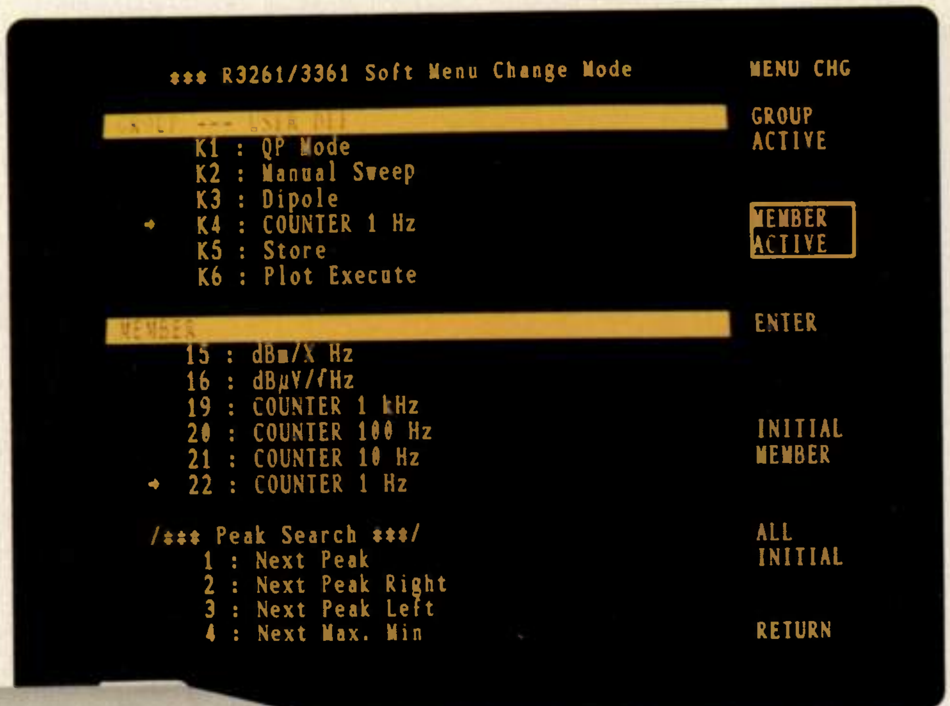

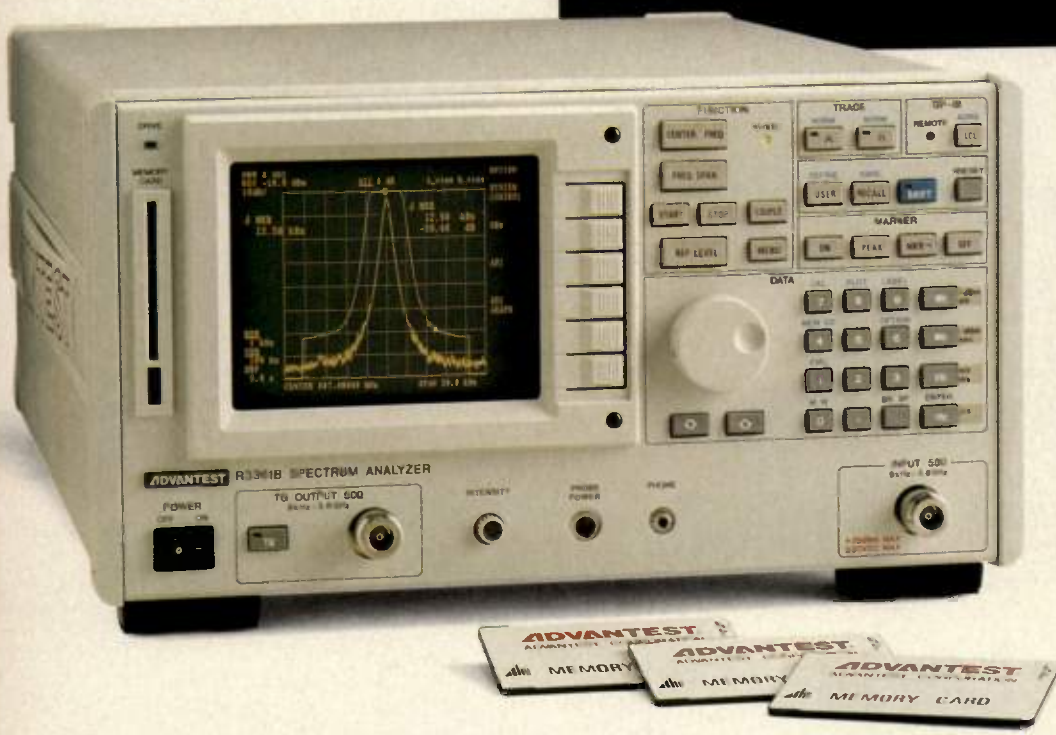

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The George Washington University

Global Positioning System: Principles and Practice
January 8-10, 1990, Colorado Springs, CO
Preparation of Signals for Digital Transmission
January 9-12, 1990, Washington, DC
Introduction to Modern Radar Technology
January 17-19, 1990, Washington, DCM
Sonar System Design and Prediction
January 29-February 2, 1990, Washington, DC
Optoelectromagnetics
February 6-8, 1990, Washington, DC
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February 12-16, 1990, Orlando, FL
Radar ECM and ECCM Systems
February 21-23, 1990, Washington, DC
Radio-Wave Propagation for Communication System Engineering
February 26-March 2, 1990, Washington, DC

Information: Misael Rodriguez, Continuing Engineering Education, George Washington University, Washington, DC 20052. Tel: (800) 424-9773; (202) 994-6106

Telecommunications Industry Association

Part 68 Rationale and Measurement Guide
January 23-24, 1990, Phoenix, AZ

Information: Suzanne Mullendore, TIA, 1722 Eye Street, N.W., Suite 440, Washington, DC 20006. Tel: (202) 457-4937

UCLA Extension

Superconductivity: Basic Concepts and Applications
January 16, 1990, Long Beach, CA
Power Electronic Circuits: Theory and Practice
January 29-February 2, 1990, Los Angeles, CA
RF and Microwave Circuit Design I: Linear Circuits
February 5-9, 1990, Los Angeles, CA
RF and Microwave Circuit Design II: Nonlinear Circuits
February 12-16, 1990, Los Angeles, CA
High-Power Microwave Sources and Applications
February 20-22, 1990, Los Angeles, CA

Information: UCLA Extension, Engineering Short Courses, 10995 Le Conte Avenue, Los Angeles, CA 90024. Tel: (213) 825-1047

University of Wisconsin-Madison

Electrical Grounding of Communication Systems
January 17-19, 1990, Madison, WI
Design of Fiber Optic Communication Systems
January 24-26, 1990, Madison, WI

Information: University of Wisconsin-Madison, College of Engineering, 432 N. Lake Street, Madison, WI 53706. Tel: (800) 262-6243; (800) 362-3020

Compact Software, Inc.

Computer-Aided Design Seminar
January 9, 1990, Bedford, MA
February 9, 1990, Salt Lake City, UT
March 6, 1990, Ottawa, Ontario, Canada
April 17, 1990, King of Prussia, PA
May 8, 1990, Bellevue, WA
June 12, 1990, College Park, MD

Information: Sam Benzacar, Training Manager, Compact Software, Inc., 483 McLean Boulevard, Corner 18th Avenue, Paterson, NJ 07504. Tel: (201) 881-1200

E/J Bloom Associates, Inc.

Modern Power Conversion Design Techniques
February 19-23, 1990, San Diego, CA

Information: Mrs. Joy Bloom, E/J Bloom Associates, Inc., Educational Division, 115 Duran Drive, San Rafael, CA 94903. Tel: (415) 492-8443

Hewlett-Packard Co.

Designing for Electromagnetic Compatibility (EMC)
December 11-12, 1989, Mountain View, CA
December 14-15, 1989, Mountain View, CA

Information: Hewlett-Packard Co., 3000 Hanover Street, Palo Alto, CA 94304. Tel: (800) 2HP-EDUC

Learning Tree International

Introduction to Datacomm and Networks
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January 23-26, 1989, Los Angeles, CA
Introduction to Fiber Optic Communications
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January 30-February 2, 1990, Montreal, Quebec, Canada
Digital Signal Processing: Techniques and Applications
January 16-19, 1990, Washington, DC
February 6-9, 1990, San Diego, CA

Information: John Valenti, Learning Tree International, 6053 W. Century Boulevard, P.O. Box 45974, Los Angeles, CA 90045-0974. Tel: (800) 421-8166; (213) 417-8888

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January 24-26, 1990, Los Angeles, CA
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January 31-February 2, 1990, Los Angeles, CA
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February 5-7, 1990, San Jose, CA
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December 13-14, 1989
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January 6-9, 1990
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ATE and Instrumentation West
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January 15-18, 1990
SMART VI Conference and Exhibition
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February 3-10, 1990
1990 IEEE Aerospace Applications Conference
Mountain Haus Hotel, Vail, CO
Information: Fausto Pasqualucci, Hughes Aircraft Corp., 3100 Fujita Street, Torrance, CA 90509.

February 13-15, 1990
IEEE Instrumentation and Measurement Technology Conference (IMTC/90)
Red Lion Inn, San Jose, CA
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February 14-16, 1990
1990 IEEE International Solid-State Circuits Conference
San Francisco Hilton, San Francisco, CA
Information: Courtesy Associates, 655 15th Street, N.W., Suite 300, Washington, DC 20005. Tel: (202) 347-5900

February 26-March 1, 1990
Electronic Imaging West 90
Pasadena Convention Center, Pasadena, CA
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March 26-29, 1990
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March 27-29, 1990
RF Technology Expo '90
Anaheim Convention Center, Anaheim, CA
Information: Kristin Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-2600; (800) 525-9154

A Look at RF Cables and Connectors

By Mark Gomez
Technical Editor

The RF cable and connector business appears to be heading for uncertain times. Even though the industry is faced with a declining military budget, new companies continue to penetrate these product lines. There are no visible signs of any cable and connector manufacturers being forced out of business. Some experts do, however, predict a shake-out sometime in the future.

"New players are popping up every-day," observes Colin Goff, sales manager at Applied Engineering Products. Walter Grant, marketing specialist at W.L. Gore Electronic Product Division, says, "I do not see an upswing in the market for a while. In fact, it is ironic in that more people are getting into the business." There have also been a few acquisitions going on in this industry. Steve Perry, senior applications engineer at Kaman Instrumentation Corporation, says, "As opposed to people going out of business, there are acquisitions taking place." This view is shared by Steve Ulett, sales manager of Penstock, Inc. "There is a consolidation of connector companies, where some firms are being bought-out by others."

"Within the cable and connector industry, there is going to be a shake-out next year. Some companies will survive, others will be merged and some be will be autonomous," predicts Ulett. Grant foresees this as a necessity. "A shake-out is going to have to happen," he says. The cable and connector business does not seem to be affected by military markets as much as other business sectors are. "The defense market is shrinking rapidly, but the cable part of it is not shrinking quite as rapidly," remarks Grant. He adds that the cable business is less volatile since the military is always retrofitting existing systems and replacing cables. "The emphasis is towards long-term programs and refits of existing programs," says Perry. Daniel Roth, manager of market development at Kaman, emphasizes that even though the market may have shrunk for the short term, Kaman's market share has not decreased and is in fact probably on the upswing.

There is a certain amount of activity in the commercial side of the business. Growing industries like cellular radio are having some impact on certain segments of the connector business. "A major opportunity exists if you could get involved in the cellular radio marketplace. It is exploding right now!" says Grant. Bob Perelman, HELIAX[R] special products marketing manager at Andrew Corporation, anticipates some growth for 1989 and points out some figures from Andrew's 1988 Annual Report. "For 1988, cellular customer demand and new products led to a 16 percent sales gain," he explains. Scott Spencer, marketing manager at M/A-COM Omni Spectra, feels there are some new opportunities available. "We do not see a lot of growth, but we do see a lot of opportunities and in particular, we see OEMs investing in new technologies to try and position themselves for the future," he reveals. The RF medical business is one that is definitely growing. "In the medical industry, for NMR applications, there is a need for high frequency cable that will not become magnetized," says Ulett.

To continue a growth curve with the level of competition out there, many companies feel that closer customer relationships are a key. "There is growth for the right companies who spend time building a quality product and work closely with a customer to learn the real application and design products accordingly," says Jim Rawlins, sales manager at Southwest Microwave.

Product Trends

As far as new trends in the cable and connector business, many companies seem to be pushing the frequencies of their product lines higher and higher. Other trends include sub-miniaturization and, to some extent, surface-mount technology. "Now, 26.5 GHz is commonplace. Forty gigahertz test sets are becoming more and more common and people are already talking about 60 GHz, and in some cases up to 90 GHz for coaxial systems," explains Paul Burgis of the RF & Microwave Division of Suhner Electronics Limited, a subsidiary of Huber + Suhner AG. There is also apparently some rejuvenated interest in a connector series that has been around for a while. "We have seen some renewed interest in the SC connector and have developed some for use with our cables," observes Perelman.

In cables and cable assemblies, there is a need for products with better phase stability. "Some of the changes I see in cable assemblies are that companies are requiring more phase stable cable and temperature compensated cables," states Ulett. Norbert Sladek, director of technology at Amphenol RF/Microwave Operations, sees a trend towards higher frequencies and better performance. "The changes I see in cables and connectors over the next few years are higher frequencies above 18 GHz, higher performance requirements meaning better VSWR and insertion losses, and probably to some degree higher capabilities for performing in adverse environments."

At RF frequencies, there is some ongoing work on coaxial contacts in a multipin connector. "There is a move towards hybrid connectors that have some coaxial pins and power pins," says Goff. Spencer says that Omni Spectra has developed a high-density coaxial multipin connector that uses their blindmate and millimeter-wave technology.

Jack Kaufman, owner and president of Aviel Electronics explains, "We are seeing more and more activity in surface-mount." But, various problems have been encountered in the pursuit of this mounting configuration. One problem is stress to the solder joint during mating. Various hold-down techniques are being studied that will not only help this problem but also hold the connector in place during soldering. Other factors such as standoff height, material considerations and terminations are being investigated as well.

All this activity in cables and connectors will no doubt benefit the RF designer. With the competition out there, he can expect to see higher levels of quality at competitive prices, and products more suitable to his needs. rf

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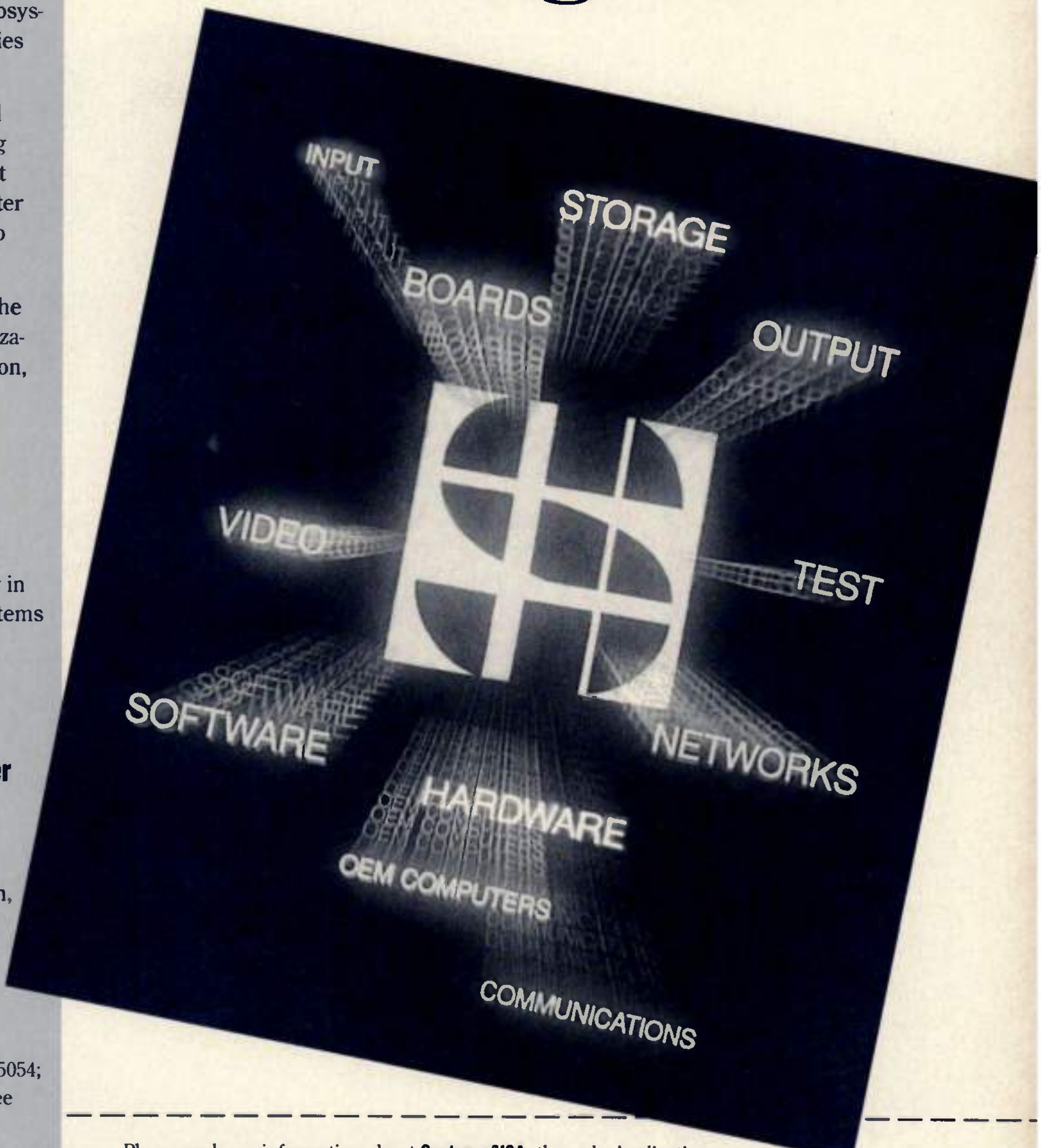

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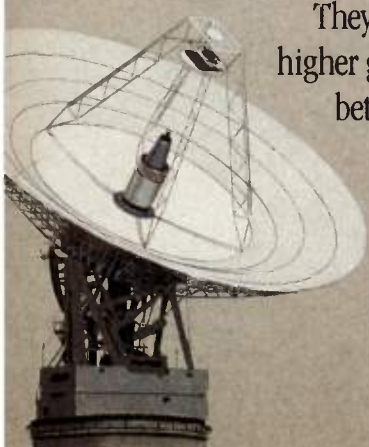

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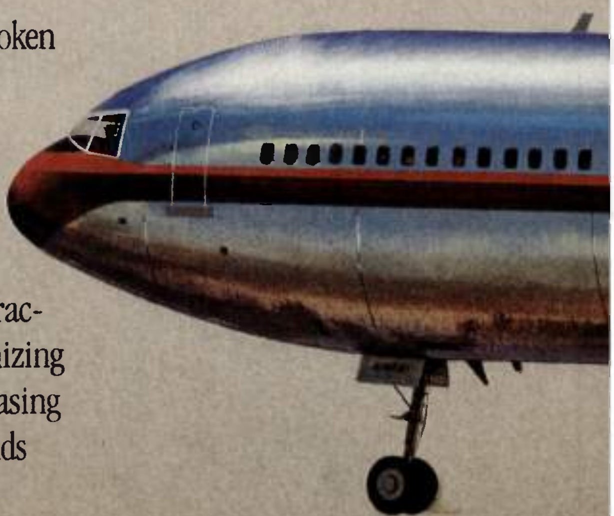

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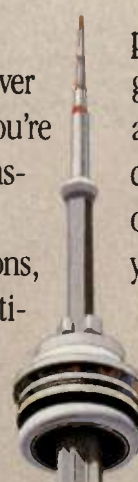

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Complex Signals: Part 1

By Noel Boutin
University of Sherbrooke

This article begins a four-part series on complex signals based on material first presented by the author at RF Expo East 88. With this first installment, the reader will acquire the ability to visualize a complex signal. Part 2 will discuss how to filter a complex signal, while Part 3 will explain how to shift in frequency a complex signal. Part 4 will present a review of some modern communication systems within which complex signals are involved.

For most engineers, the word "complex" refers to frequency domain, and, more particularly, to the Laplace complex frequency $S = \sigma + j\omega$. For some, that word also evokes something imaginary and, by consequence, not related at all to the reality. Many practitioners have thus come to regard these matters merely as academic considerations having no practical interest. This series of articles is aimed at members of this group. It is important to understand that complex signals really exist and are perhaps more frequent in modern telecommunication systems than purely real signals.

From Real to Complex Signals

Most engineers are very familiar with two-dimensional signals, such as those present at any point in a circuit or at the output connector of an apparatus. This is the kind of planar (two-dimensional) signal which can easily be seen on the face plate of an oscilloscope, where the horizontal and vertical axes represent, respectively, the time and the amplitude of the signal. Sine, cosine and square waves are among the most well-known planar signals. Since, for each time value t, the amplitude of such signals can be characterized by a real number, these signals are qualified as "real signals." That appellation is somewhat misleading because it falsely implies the existence of invisible "imaginary signals." A more meaningful nomenclature for real signals would be "two-dimensional signals."

Open-minded engineers can easily accept that signals are not compulsorily forced to lie on a plane, i.e., to be two-dimensional signals. Mechanical engineers, for one, currently use excitations whose evolutions with time cannot be represented by a two-dimensional signal. Consider, for example, the eccentric rod fixed to the spinning wheel shown in Figure 1. To simplify the discussion and present a fundamental complex signal, assume that the wheel is spinning at a constant angular speed. The time evolution of the rod position in the x-y plane can then be represented by the three-dimensional signal c(t) shown in Figure 2. For each time value t, two *real* numbers, x and y, are required to fully position the amplitude of that three-dimensional signal in the x-y plane. The resultant two-dimensional signals x(t) and y(t) evolve on two distinct planes perpendicular to each other. As shown in Figure 3, the signals x(t) and y(t) constitute the orthogonal projections of the three-dimensional signal c(t). Taken together, those two signals represent c(t) without any ambiguity. The three-dimensional signal c(t) is called a "complex signal," and can be represented mathematically as the vectorial sum of the two orthogonal *real* signals x(t) and y(t), or

$$c(t) = x(t) + jy(t) \tag{1}$$

Here, x(t) is the "real" part of c(t), and jy(t) is the "imaginary" part of c(t). The letter j represents $\sqrt{-1}$, the "imaginary" unit number. The word "imaginary" does not mean that y(t) is not physically realizable. It only means that the real signal y(t) lies on a plane perpendicular to the plane on which x(t) evolves. This implies that a complex signal c(t) cannot be transmitted through a conventional transmission line like a coaxial cable or a microstrip. Indeed, such lines can only support simultaneously real signals lying on the same plane. The solution lies in using two transmission lines to vehiculate the complex signal c(t), each one carrying one of the two real signals x(t) and y(t), the orthogonal projections of c(t). In effect, this is what is done very often in modern telecommunication systems. More about that will be said in future installments.

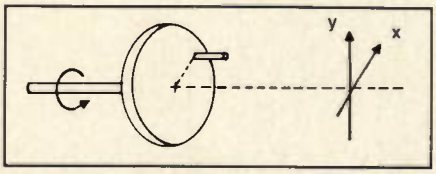

Figure 1. A spinning wheel with eccentric rod.

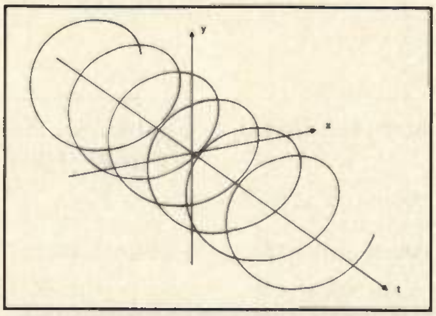

Figure 2. Time evolution of the rod position in the x-y plane of the constant speed wheel shown in Figure 1.

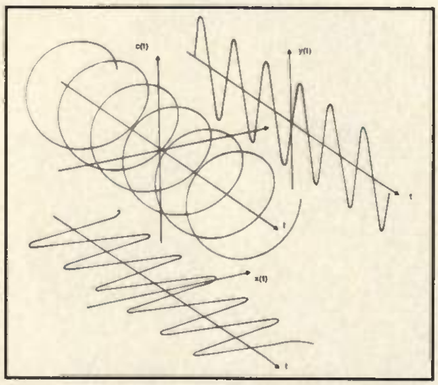

Figure 3. Decomposition of the complex signal c(t) into its two orthogonal projections x(t) and y(t).

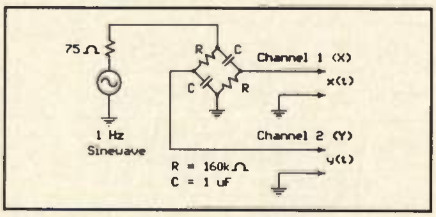

Figure 4. The synthesis of the complex signal $e^{j\omega t+\theta}$

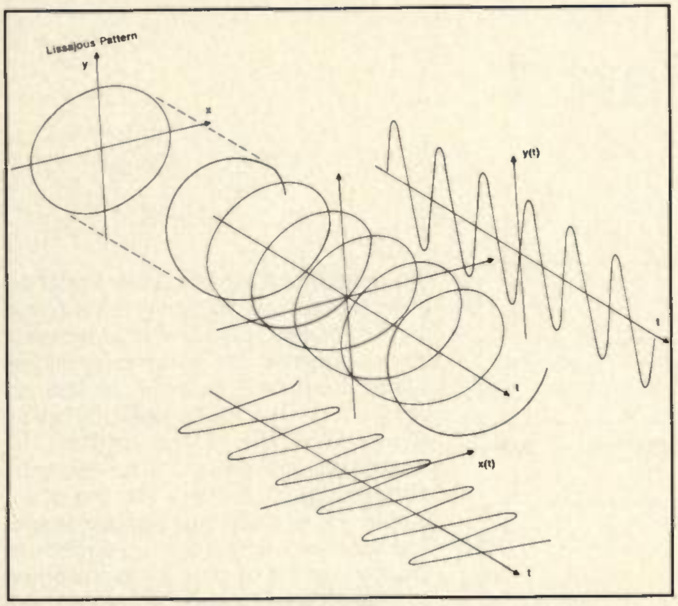

Figure 5. The relationship between a complex signal and its Lissajous pattern.

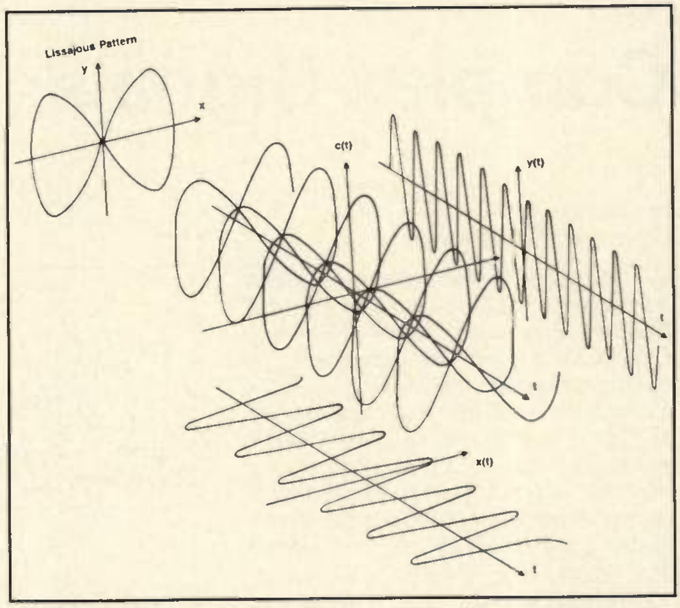

Figure 6. Example of a non-constant envelope complex signal.

According to Figure 3 and the relation given in equation 1, the particular complex signal c(t) so illustrated is given by:

$$c(t) = \cos \omega t + j \sin \omega t \quad (2)$$

which, according to the Euler relation,

$$e^{j\theta} = \cos \theta + j \sin \theta \quad (3)$$

can also be expressed as:

$$c(t) = e^{j\omega t} \quad (4)$$

This complex signal is known as the exponential, a member of the most basic signal family of complex exponential e^{st}. The parameter ω is the angular speed of rotation in radians per second of the exponential. The exponential is not constrained to rotate only in the direction shown in Figure 3. It can also rotate in the reverse direction. In that case, the angular speed of rotation (or angular frequency, as it is usually called) is negative and the exponential is expressed as:

$$e^{-j\omega t} \quad (5)$$

The sign of the angular frequency ω is therefore representative of the *direction* of rotation. Neglecting this basic discussion about the angular frequency of an exponential signal leads to the most current fallacy in electrical engineering: Negative frequencies do not exist; they are just a mathematical artifice. The main reason why this fallacy becomes established is that, with one exception (1), the exponential shown in Figure 3 is never pictured in modern textbooks on signals, circuits and systems. Students are accustomed to seeing and working only with the orthogonal projections of the exponential, i.e., the cosine and the sine waves. As:

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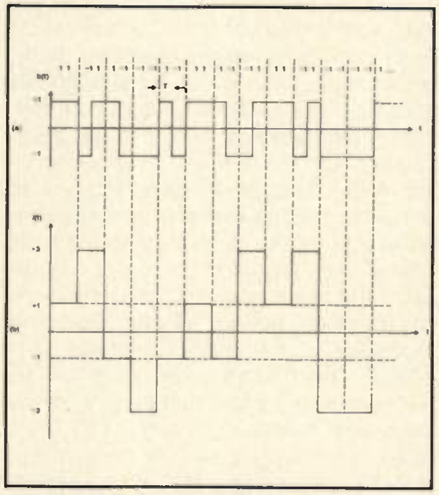

Figure 7. (a) An NRZ binary signal b(t), and (b) a four-level Gray-encoded version of b(t).

$$e^{-j\omega t} = \cos(-\omega t) + j \sin(-\omega t) \tag{6}$$

and

$$\cos(-\omega t) = \cos(\omega t) \tag{7}$$
$$\sin(-\omega t) = -\sin(\omega t),$$

it is natural but wrong to defend the non-existence of negative frequencies.

The Lissajous Pattern

For those not yet convinced of the existence of negative frequencies, consider the laboratory set-up shown in Figure 4. A sinewave of about 1 Hz is connected to a phase-shifting bridge. The highpass arm of the bridge introduces a +45 degree phase shift, while the lowpass one introduces a −45 degree phase-shift. The two resultant signals can be thought of as orthogonal projections of the complex signal $e^{j\omega t + \theta}$, since:

$$x(t) \equiv \cos(\omega + \theta) \tag{8}$$
$$y(t) \equiv \sin(\omega t + \theta)$$

where θ is an immaterial unknown constant phase angle. Those two signals are then connected to an oscilloscope via channel 1 (X) and channel 2 (Y). Setting the amplitude switches of both channels to the same value and the time base switch of the oscilloscope to mode X-Y (no internal sweep), one can observe a circular trace rotating anticlockwise at a constant speed of 1 revolution/sec. Known as a Lissajous pattern (2), this trace is nothing but the projection on the x-y plane of the exponential $e^{j\omega t + \theta}$. This is illustrated in Figure 5. Pressing the channel 2 (Y) invert switch of the oscilloscope leads, according to the relations given in equations 6 and 7, to a circular trace rotating in the reverse direction, i.e., to the projection of the exponential $e^{-j\omega t + \theta}$.

Although the Lissajous pattern has been known for a long time, it appears never to have been used to visualize the direction of evolution of a complex signal and, in particular, the direction of rotation of an exponential. Incidentally, the same experiment run at 100 Hz instead of at 1 Hz (160 kohms → 1.6 kohms) would only display permanent circles on the oscilloscope due to persistence of the CRT phosphor, and the response of the human eye. With this display, one could measure the relative phase between the input sine waves, but it would not help anyone to discover the existence of two possible directions of rotation (and negative frequencies).

So far, only one particular complex signal, the exponential $e^{j\omega t}$, has been discussed and illustrated. That complex signal is special in the sense that its module remains constant with time. This leads to a Lissajous pattern which is a circle. In general, complex signals can evolve in any way, thus generating all kinds of Lissajous patterns. As an example, Figure 6 illustrates the butterfly-like Lissajous pattern corresponding to the following complex signal:

$$c(t) = \cos \omega t + j \sin 2\omega t \tag{9}$$

Until now, only complex signals whose orthogonal projections are sinewaves have been illustrated. However, any other waveforms, periodic or not, are also admissible (see next section).

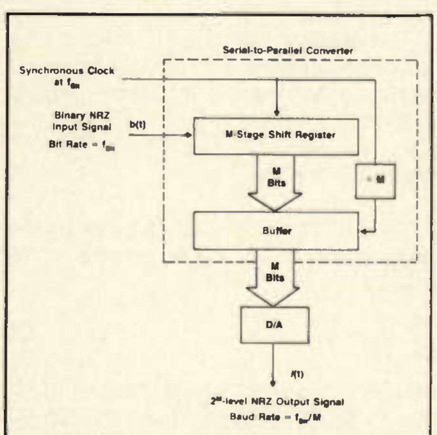

Figure 8. The conversion of a binary NRZ signal into a 2^M–level NRZ signal.

Complex Signals in Data Transmission

Consider the binary signal b(t) illustrated in Figure 7(a). Each bit of duration T is formatted in bipolar Non-Return-to-Zero (NRZ-format). The bit rate f_{Bit} equals (1/T) bits/sec. It is possible to reduce that bit rate by a factor M (and by consequence, reduce the occupied bandwidth) by transforming the binary signal b(t) into a 2^M-level signal $l(t)$. This can easily be done, as shown in Figure 8, with an M-bit serial-to-parallel converter followed by an M-bit digital-to-analog converter (D/A). The original binary signal b(t) is segmented in successive blocks known as baud, each one comprising M bits. To each specific baud of duration MT corresponds a particular amplitude level of $l(t)$. In order that, at the receiver, baud errors result in the lowest possible number of bit errors, it is interesting to choose an encoding rule such that adjacent amplitude levels differ by only one bit. Such an encoding is known as a Gray encoding. As an example, Figure 7(b) shows a four-level Gray-encoded signal corresponding to the binary signal b(t) of Figure 7(a). Note that in order to preserve the noise immunity, the amplitude of $l(t)$ must be set higher than the amplitude of the binary signal b(t). This is the price that must be paid for the bandwidth reduction obtained with the multi-level solution. Assuming all the l_i levels are equally likely, the mean power p of the L-level signal $l(t)$ is given by:

$$\bar{p} = \frac{1}{L}\sum_{i=1}^{L} (l_i)^2 = \frac{L^2 - 1}{3} \qquad (10)$$

where $L = 2^M$.

A four-level signal therefore requires five times (about 7 dB) more power than a binary signal. This is quite a high price to pay for a bandwidth reduction by half.

Complex Signals as Power Savers

It is possible to reduce the transmitted power without affecting the noise immunity if one is willing to transform the real L-level signal $l(t)$ into a complex signal c(t). One such nonlinear transformation is shown in Figure 9. Here, the upper and lower levels of the four-level signal shown in Figure 7(b) have been bent 90 degrees forward with respect to their original positions. The resultant U-shaped complex signal c(t) evolves at the surface of a square parallelepiped whose edges coincide with each one of the four levels of the original four-level real signal $l(t)$.

For reasons that will be detailed in a future installment, it is the usual practice in data transmission to denote c(t) as follows:

$$c(t) = I(t) + jQ(t) \qquad (11)$$

The mean power p of c(t) is given by the summation of the mean powers of I(t) and Q(t), or:

$$\bar{p} = \bar{p}_I + \bar{p}_Q \qquad (12)$$

For the complex signal illustrated in Figure 9, I(t) and Q(t) are both binary signals. That complex signal therefore requires only twice the power of the original binary signal as opposed to five times more with $l(t)$.

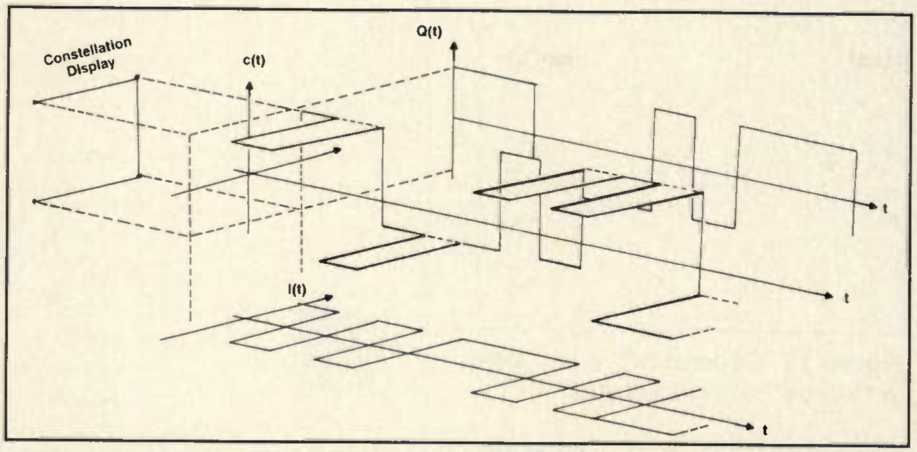

Figure 9. The generation of a four-state complex signal c(t) by folding the four-level real signal /(t) of Figure 7(b).

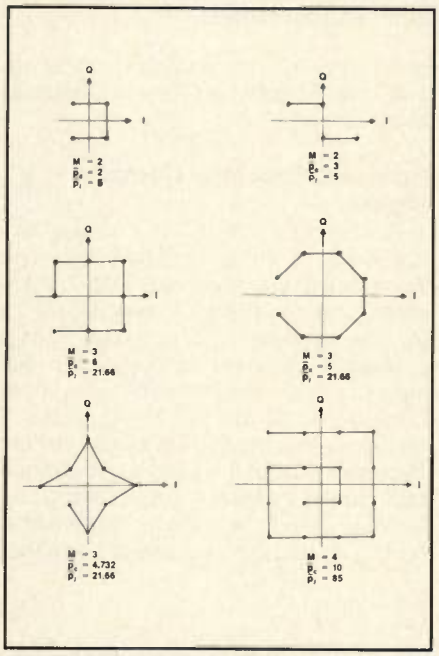

Figure 10. Constellation displays of folded four-, eight- and 16-level signal /(t).

The Constellation Display

In data transmission, the Lissajous pattern is known as the constellation display or signal-space diagram. Figure 9 shows such a constellation display for the four-state complex signal discussed earlier. Due to the inherent persistence of the CRT, each one of the four states appears as a bright spot, and each rapid transition from one state to another as a light link between those spots.

The folding operation can be applied to virtually any L-level real signal /(t), irrespective of its number of discrete levels. Furthermore, there is, in theory, no restriction on the shape of the resultant complex signal and its accompanying constellation display except that the receiver must be able to unfold the received complex signal and restore /(t). Of course, it is advantageous to use a folding technique which leads to a low-power, narrow-bandwidth, complex signal which can easily be unfolded at the receiver!

Figure 10 shows some possible constellation displays resulting from the folding of a four-level, an eight-level, and a 16-level signal /(t). In order to illustrate the power gain which can be realized, each constellation is accompanied by the mean power value $\bar{p}_c$ of the corresponding complex signal c(t), and by the mean power value $\bar{P}l$ of the L-level signal /(t) before folding. Both of these values have been normalized with respect to the original binary signal mean power. All of these constellations are open in the sense that there is no direct link between the lower and upper state levels of /(t). Incidentally, it is that characteristic which permits unfolding of the received signal in an unambiguous way.

Complex Signals as Bandwidth Savers

The bandwidth occupancy of the complex signals c(t) corresponding to the constellations illustrated in Figure 10 increases with L, since the signal is not allowed to transit straight through the shortest path from one state to another. Rather, the signal must travel in a backward and forward motion, following the only allowable path linking all the states. Instead of folding the L-level real signal /(t) in the manner discussed so far, it is possible to transform it into another complex signal leading to the same states as those

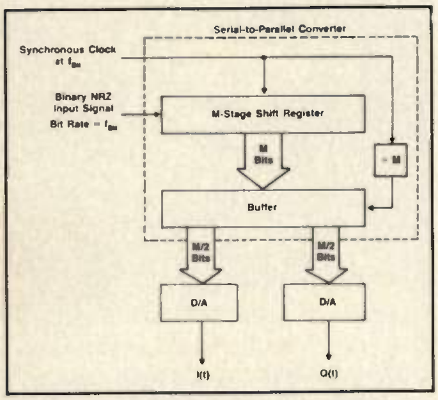

Figure 11. Generation of the complex signal $c(t) = I(t) + jQ(t)$.

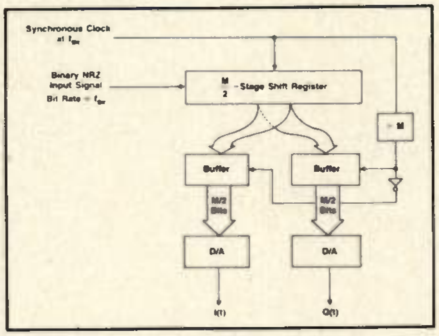

Figure 12. Generation of an "offset keyed" complex signal.

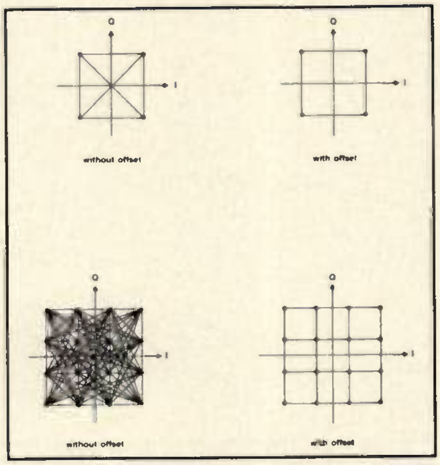

Figure 13. Constellation displays of complex signals generated with and without offset.

illustrated in Figure 10. This time, however, the signal is allowed to transit in a direct line from one state to any other one. For the case when M is even, such a complex signal can easily be generated, as shown in Figure 11. The use of linear D/A converters leads to square constellation displays which present the same noise immunity as the previous square ones. Indeed, in such cases, the full distances between each state is preserved in signals I(t) and Q(t), from which the receiver now makes its decisions. Any constellation link non-orthogonal to the I-Q axis would lead to shorter distances between the corresponding states of the I(t) and Q(t) signals and, accordingly, to a decrease in noise immunity. The bandwidth of the complex signal generated in the manner shown in Figure 11 is the same as the one of the real L-level signal $I(t)$ generated as shown in Figure 8. There is, however, as Figure 10 illustrates, a net difference between the power of each signal required to achieve the same noise immunity. That explains the present popularity of such complex signals over real signals in modern power- and bandwidth-efficient digital communication systems.

Offset Keying

Complex signals generated in the manner illustrated in Figure 11 present larger amplitude fluctuations than with the previous method, due to the direct transitions from one state to another. The transmission of such complex signals through nonlinear amplifiers generally results in poorer performances, due to self-generated interference between I(t) and Q(t), the two orthogonal projections (3). Amplitude fluctuations can be reduced to their minimum by forbidding simultaneous renewals of the I(t) and Q(t) buffer contents. This can easily be done, as shown in Figure 12, by clocking the buffers with two synchronized clocks, one being offset by M/2 bits or half a baud with respect to the other. In that way, the complex signal can still transit from one state to any other one, but in two successive orthogonal moves instead of diagonally. The resultant constellation displays for M = 2 and M = 4 appear in Figure 13, along with those obtained without offset. This kind of signalling is known as "offset keying". It is worth nothing that the bandwidth occupancy of the complex signal remains the same whether or not an offset is introduced.

Constant Envelope Complex Signals

In situations where the complex signal needs a more constant envelope when transiting from one state to another, one can lay the L-level signal $I(t)$ on the surface of a constant radius cylinder instead of folding it as discussed earlier. The radius of the cylinder is chosen such that the $2^M = L$ levels of $I(t)$ are evenly distributed all around the circumference of the cylinder. The resultant complex signal c(t) of power $(L/\pi)^2$ is a constant amplitude exponential having up to L discrete phase positions:

$$c(t) = (L/\pi)\, e^{jI(t)} \qquad (13)$$

The corresponding constellation displays for M = 2, 3 and 4 appear in Figure 14, along with the power p_c of the complex signal c(t) given in equation 13. As can be seen, the required power increases with M much more rapidly with the exponential signals than with the complex signals leading to the square constellations discussed earlier. Furthermore, to take full advantage of the processing done on $I(t)$ at the transmitter, the receiver must unroll the received complex signal instead of processing independently its orthogonal projections I(t) and Q(t). This is mandatory if the same noise immunity is sought. In fact, from a strict power point of view, only the case M = 2 presents some interest. Appropriately processed at the receiver, that constant envelope four-phase complex signal requires 1.2337 (approximately 0.9 dB) less power than the corresponding square four-state complex signal in order to achieve the same noise immunity. Of course, all the constant envelope complex signals discussed so far still require less power than their real counterpart signals $I(t)$. This is because, for $L \geq 2$, the following power inequality is always verified:

$$\frac{L^2}{\pi^2} < \frac{L^2 - 1}{3} \qquad (14)$$

From Reference 4, the bandwidth occupancy of the constant envelope complex signal given in equation 13 is about the same as the one of $I(t)$. Essentially, this comes from the fact that the nonlinear processing done on $I(t)$ is not very severe. As can be seen from the constellation displays, irrespective of the value of L, $I(t)$ never wraps completely the cylinder surface on which it is laid. This is because, from equation 13, the radius of the cylinder increases with L.

In situations where it is possible to trade off power for bandwidth, one can fix the radius of the cylinder and allow $I(t)$ to wrap the cylinder surface more than one time. As an example, let the complex signal be given by:

$$c(t) = (4/\pi)e^{jI(t)} \qquad (15)$$

For $L > 4$, the constellation displays associated with that family of constant mean power complex signals are all identical because, as $I(t)$ travels around the cylinder, its L levels overlap and result in only four distinct states.

Needless to say, it is mandatory to unwrap the received signal in order to discriminate the L transmitted levels of $I(t)$. The bandwidth occupancy of such constant mean power complex signals grows as L increases, because the speed of rotation of the exponential doubles for each two-fold increase in L value, whereas the baud period of $I(t)$ increases only by a factor M with $L = 2^M$.

Conclusion

The complex baseband signals discussed in this first installment are not very often used "as is" in communication systems. Instantaneous transitions are physically impossible to realize and, in any case, not at all desirable. Narrower bandwidth occupancies can be obtained by filtering such complex signals before their transmission. This subject will be discussed in Part 2 of this series. The readers may also question the feasibility of transmitting such com-

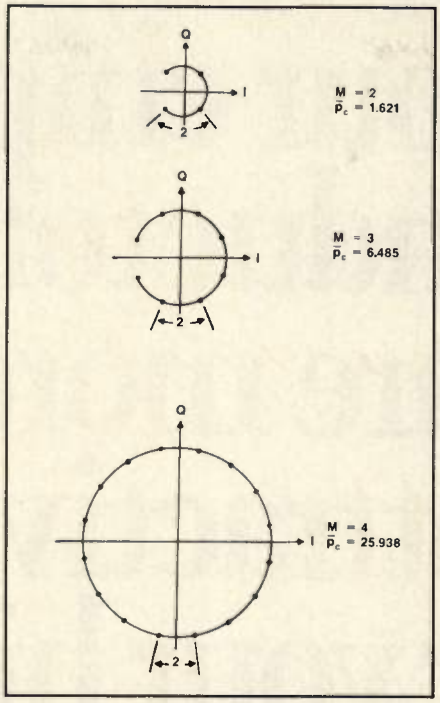

Figure 14. Constellation displays of the complex signal $c(t) = (L/\pi)e^{jI(t)}$ where $I(t)$ is a 2^M-level real signal with M= 2, 3 and 4.

plex signals over existing communication channels. That important and practical subject will be covered in the third installment. Those who still entertain some suspicions about the frequent use of complex signals in modern communication systems will have to wait until the final installment or...enjoy the pleasure of discovering this for themselves before that disclosure. rf

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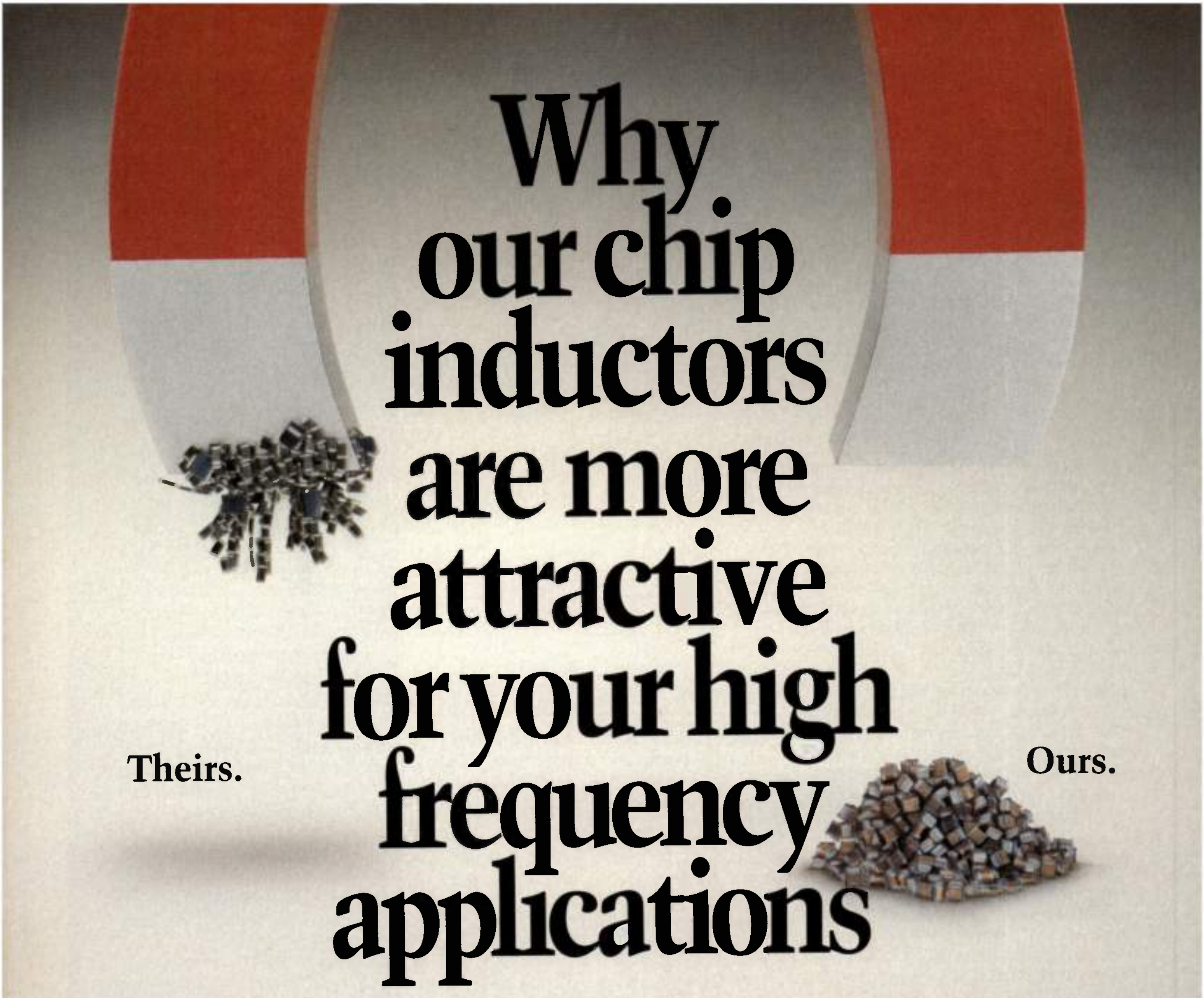

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rf design awards

A Balanced RF Oscillator

By Bránislav Petrovic
General Instrument Corp.

A signal that is symmetrical with respect to common-ground (balanced signal) is characterized by equal amplitudes and a 180 degree phase relationship. In many applications, such as in balanced mixers, phase detectors, frequency doublers, differential frequency dividers, etc., the availability of balanced oscillator (BO) signals is important, if not essential. A runner-up in the Fourth Annual RF Design Awards Contest, this article presents a balanced RF oscillator with two symmetrical outputs of equal amplitude and 180 degrees phase shift.

Normally, a balanced local oscillator (LO) signal for RF applications is derived from a single-ended (asymmetrical) LO output, by means of a balun (balanced-to-unbalanced transformer). A balun, depending on the operating frequency range, can take different forms (ferrite transformer, transmission line transformer-wire, microstrip or stripline). Usually, drawbacks such as poor balance, matching, excessive loss, limited frequency range, LO radiation (EMI), cost, size, etc., are associated with the use of baluns.

The balanced RF oscillator described here has two symmetrical outputs of equal amplitude and 180 degrees of phase shift (0 degrees and 180 degrees). The inherent phase and amplitude matching makes this oscillator particularly useful. The balanced nature of the oscillator eliminates the need for baluns, and hence, their associated problems.

The Balanced Oscillator

This oscillator has a voltage-controlled oscillator (VCO) operating in the 1.3 GHz to 1.9 GHz frequency range. The circuit schematic and parts list are shown in Figure 1. The circuit consists of two identical negative impedance oscillators, sharing a common tank circuit C_5 - Z - C_7 (C_5, C_7 are varicap diodes, Z is microstrip printed line, having the role of an inductance).

A BO is composed of two negative impedance oscillators, connected back to back, as shown in Figure 2. The two oscillators share a common tank circuit 2C - L/2 - L/2 - 2C. Therefore, both oscillators will oscillate at the same frequency, determined by the tank circuit ($f_0 = 1/(2\pi\sqrt{LC}) = 1/(2\pi\sqrt{(L/2)2c})$. The RF current I flowing in the tank circuit is common for both transistors, but has opposite signs with respect to Q_1 and Q_2. (While I is flowing into Q_1, it is flowing out of Q_2.) Assuming perfect symmetry in the circuit, voltages at emitters of Q_1 and Q_2 will have equal amplitudes and opposite signs, i.e. the signals will be perfectly balanced. At the center of symmetry (point Z), a virtual ground will be formed at the fundamental frequency f_0. The RF voltage at f_0 at this point will be zero. Therefore, the BO can conveniently be represented in equivalent form, as shown in Figure 3. From the point of view of external loads, it will look like a truly balanced generator, with mid-point grounded at RF (f_0).

The tank circuit 2C - L/2 - L/2 - 2C in Figure 2 could be replaced by a single L-C circuit, without any effect on the operation of the circuit. The only difference would be that the point of symmetry Z would not physically be accessible.

Design and Performance Considerations

The main consideration in BO performance, apart from output frequency and level, is the amplitude and phase balance of the two outputs. Depending on the application, one or both of the two parameters may be of primary importance. Typically, in applications such as mixers in frequency converters, amplitude balance is more important than the phase balance, since the LO feedthrough is far more sensitive to the amplitude imbalance than to the phase imbalance. In the case of the phase lock loop application, the phase balance is more important, since it will directly contribute to the phase error at the output of the phase detector. As an example, it can be calculated that the residual signal, when the two balanced signals are combined (summed), will be only 18 dB suppressed in the case of 1 dB p-p amplitude imbalance. For the same amount of suppression, it would take as much as 7 degrees of phase imbalance.

The BO has inherently good phase and amplitude balance. The physical symmetry of the circuit is very important, particularly with respect to the phase balance. Any differences in physical length at the two outputs of the BO will directly cause an error in phase. As an

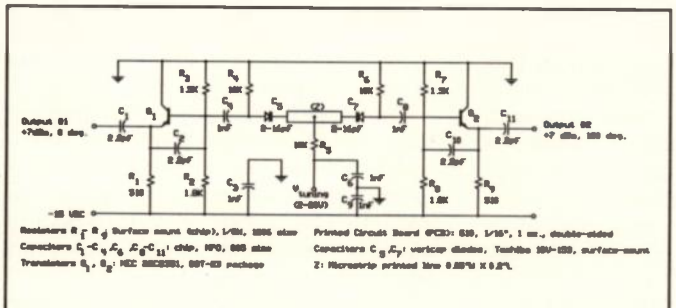

Figure 1. The 1.3 to 1.9 GHz voltage-controlled balanced oscillator.

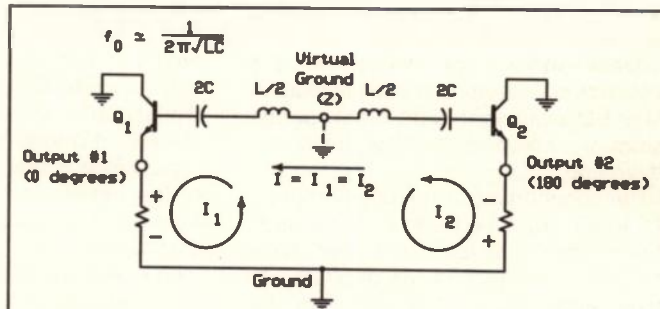

Figure 2. Simplified schematic diagram for a balanced oscillator.

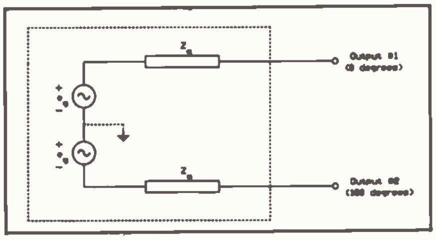

Figure 3. Equivalent circuit for a balanced oscillator.

example, on G-10 board (eff. $\varepsilon = 3.3$), 0.1 in. of line length represents 8 degrees of electrical length at $f0 = 1.5$ GHz.

Since the BO has twice the output power relative to a comparable single-ended oscillator, resistive pads can be afforded at each output, to further improve the balance. Practically, there is about a 5 dB power advantage over a standard oscillator. To obtain two balanced outputs from a single-ended oscillator, a splitter with a 3 dB split loss and typically insertion loss of 2 dB is required.

The addition of resistive attenuators at the output will also improve the BO isolation from the load, which is another important design consideration. One such consideration is a so-called "load pulling" effect, where the load can pull the frequency of the oscillator. This is particularly important when the load is varying. Another important situation is when the load has signals from other sources (e.g. in frequency converters). In this case, poor isolation can lead to frequency modulation, or even injection locking of the oscillator by external signals. However, in almost all applications, external signals will appear at both oscillator terminals, as common mode signals, and will be rejected as such. This is evident by examination of Figure 1. Namely, if two identical signals are applied simultaneously to terminals 1 and 2 from the load, no current will be produced in the base circuit, since both bases will experience identical (common mode) voltage raise. External currents will have ground return via collector terminals and will not interfere with the tank circuit. Therefore, the load isolation of the BO is superior to the single-ended oscillator, which is another important advantage.

The common mode rejection property of the BO, combined with resistive pads at the output, provides very good isolation and impedance of the oscillator. In many applications, the BO can be connected directly to the switching diodes in a mixer, eliminating the need for buffer amplifiers.

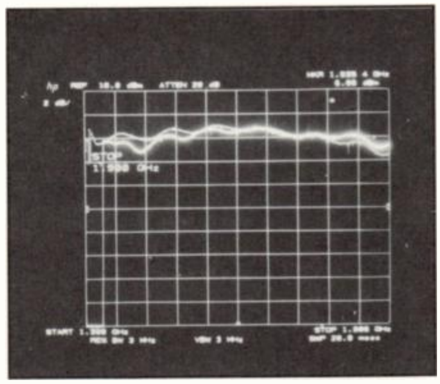

Figure 4. Output levels of the balanced oscillator; trace 1: output 1, trace 2: output 2.

Another important consideration in oscillator design is frequency stability and/or phase noise. As far as frequency stability is concerned, there is no inherent advantage of a BO over a single-ended oscillator. It is worth mentioning that any frequency stabilizing element, such as a crystal or SAW resonator, can be used in place of a tank circuit in a BO, utilizing the series resonance properties of the element.

In the case of phase noise, a BO apparently has an advantage over a single-ended oscillator. Namely, the phase noise components generated in each of the transistors in a BO, being random in nature and thus uncorrelated, will add up on a power basis (3 dB), whereas the LO signals of each of the transistors, being coherent, will add up on a voltage basis (6 dB). This will yield a net improvement of 3 dB over a single-ended oscillator. However, the phase noise is partially caused by the losses in the tank circuit itself, acting as a common phase noise. Due to this noise, the overall effect will be somewhat less than 3 dB.

Content of harmonic signals is also an important consideration. Low harmonic content is usually desired, due to potential problems in the load circuitry (e.g. spurious responses, saturation, etc.) and also due to radiation problems (EMI). Due to the balanced nature of the signals, all odd-order harmonics in a BO will be greatly reduced. Of particular importance is the leakage and radiation of the fundamental signal f_0. The leakage to power supply lines and other surrounding circuitry of the fundamental signal and odd harmonics is considerably reduced in a BO. For even-order harmonics, similar characteristics as with a single-ended oscillator could be expected.

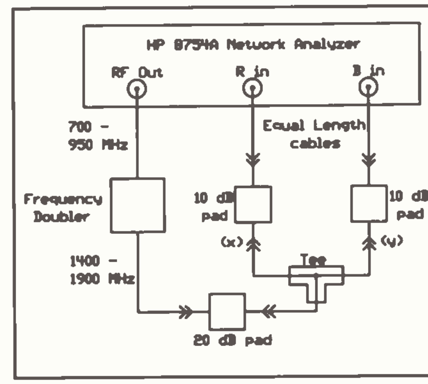

Figure 5(a). Zero phase reference calibration.

Finally, it is worth mentioning that there are no inherent limitations in the frequency of operation of a BO, other than those which apply to any other oscillator.

Balanced Oscillator Measured Performance

A voltage-controlled balanced oscillator was designed and built, per Figure 1 schematics and parts list. The oscillator operates from 1.3 GHz to 1.9 GHz. It tunes to 1260 MHz at 0 V, and to 1940 MHz at 24 V tuning voltage. Several parameters have been measured, including output power, amplitude and phase balance, and harmonic content. The measurements are discussed below.

As can be seen in Figure 4, the power output at each port (1 and 2), is about +6 dBm at both low and high band, and +8 dBm in midband. (1 dB test cable loss has been added to the readings.) The amplitude balance is almost perfect in the midband (0 dB), and falls off towards low and high band, to 1 dB p-p. The small ripple in the response in Figure 4 is due to a mismatch of the BO and cable/spectrum analyzer. If a pad (10 dB or so) is added to each output of the BO, the ripple would disappear and the amplitude balance will be better than 0.5 dB p-p. Of course, while one port is being measured, the other port of the BO must be terminated into 50 ohms.

The phase balance measurement took special precautions to obtain accurate measurement. A special setup had to be devised, as shown in Figures 5(a) and 5(b). First, a calibration for 0 degree phase reference was done, by using two equal length cables connected via 10 dB pads to a tee, as shown in Figure 5(a). A tee was used (rather than a splitter) to minimize phase error. (Phase balance uncertainty of a splitter could

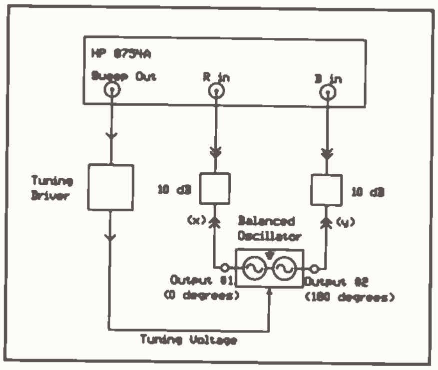

Figure 5(b). Phase balance measurement set up for the balanced oscillator.

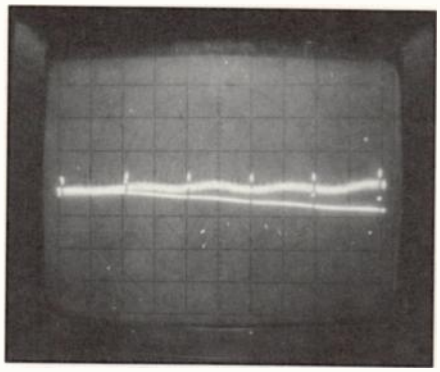

Figure 6. Phase balance of the balanced oscillator (upper trace: zero phase reference; lower trace: phase of output #1 relative to phase of output #2).

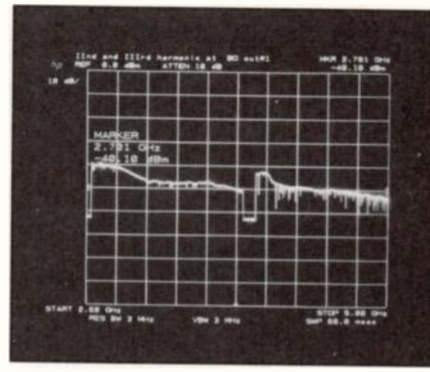

Figure 7. Balanced oscillator harmonic levels (left half trace: 2nd harmonic; right half trace: 3rd harmonic).

cause an error.) Next, the BO was connected to points (x) and (y), as shown in Figure 5(b). The tuning driver was adjusted for BO sweep of 1.4 to 1.9 GHz. Since the tuning linearity of the BO vs. tuning voltage was fairly good, the sweep linearity of the BO for this measurement was acceptable. The phase response of the BO is shown in Figure 6. A total of 8 degrees phase unbalance across the 500 MHz range was measured. At 1.4 GHz the phase matching is almost perfect (180 degrees), and the phase matching linearly degrades, to reach 8 degrees unbalance at 1.9 GHz. This almost linear phase, indicates that a fixed line length difference is in question. This could be caused by a small asymmetry in the circuit line lengths, or by a residual measurement error (e.g., after reconnection of test cable connectors/pads from calibration set up to the BO ports). Also a fixed phase error could occur while changing the network analyzer reference line from 0 degrees to 180 degrees (i.e., instrument uncertainty). So, the overall 8 degrees of phase unbalance can be taken conservatively.

Finally, the harmonic content of the BO output was measured. The result is shown in Figure 7 (second and third harmonic levels, when fundamental swept from 1.3 to 1.9 GHz). The actual levels are 10 dB higher than the readings in Figure 7, due to a 10 dB external pad used. The highest is the second harmonic, −30 dBm at 2.7 GHz. This represents 37 dBc suppression below the carrier. The third harmonic is better than 40 dBc.

Overall, the BO demonstrated excellent performance, which could hardly be matched or surpassed by any single-end oscillator/balun combination, particularly in this frequency range. The BO (Figure 1) measured performance is:

- tuning range: 1.3 GHz to 1.9 GHz
- output power (each output): +7 dBm ±1 dB
- amplitude balance: ±0.5 dB
- phase balance: ±4 degrees
- harmonic suppression: 37 dBc

Balanced Oscillator Applications

Balanced oscillators are very suitable in balanced mixers (single or double balanced), phase lock loops, synthesizers, and generally anywhere where a 0/180 degree biphase LO signal is needed. Its function is equivalent to a single-ended/balun combination. Furthermore, it can directly provide differential drive for differential amplifiers, such as in frequency dividers and prescalers. Also, if properly driven it can function as an injection-locked oscillator (ILO).

The most typical application would be in balanced mixers, and hence in a complete range of applications (including modulators/demodulators, up/down converters and many others). To demonstrate suitability of a BO, a double balanced mixer was built, using the prototype BO of Figure 1.

Balanced Oscillator in a Double Balanced Mixer Circuit

A double balanced mixer with the BO is shown in Figure 8. As can be seen from the schematic, it is similar to the regular double balanced mixer (DBM). The only difference is that one transformer is replaced by a BO. Internal RF, IF and LO currents are identical as in a regular DBM. During half a BO cycle, one diode pair is turned on (e.g., D_1 and D_2), while in another BO cycle, diodes D_3 and D_4 are turned on. The RF and IF currents will flow into both BO ports, as common mode signals. Therefore, the BO has to provide ground return for these currents. So, a low impedance at RF and IF frequencies of the BO is required. RF and IF currents will flow partially through shunt branches of attenuator pads to ground, and partially through transistors in BO to ground. The RF to BO and IF to BO isolation will depend on attenuation and symmetry of attenuator pads and on impedance balance of the BO, as well as on diodes and transformer balance and symmetry. The same applies for other port-to-port isolation (BO to RF and IF, and RF to IF isolation). Given sufficient BO level, the insertion loss and compression point will depend primarily on transformer and diode characteristics.

The mixer of Figure 8 was tested at IF = 400 MHz (RF = 1.0 to 1.5 GHz; BO = 1.4 to 1.9 GHz). The mixer showed very good isolation and compression characteristics. However, the insertion loss appeared to be too high (12 dB). It was later traced to excessive insertion loss of a ferrite transformer (5-6 dB) at the RF frequency. (For lack of higher frequency core, this core, which is normally used up to 600 MHz, had to be used). This was also the reason for an unusually high 1 dB compression point, measured in excess of +9 dBm, since the high balun insertion loss reduced the effective RF level applied to the switching diodes. The performance of the balanced oscillator/double balanced mixer is shown below:

- BO frequency range: 1.4 to 1.9 GHz
- RF frequency range: 1.0 to 1.5 GHz
- IF frequency: 400 MHz
- Conversion loss: 12 dB
- BO level at IF port: −20 dBm
- BO second harmonic at IF: −22 dBm
- BO level at RF port: −19 dBm
- 1 dB compression point: +9.7 dBm at RF in

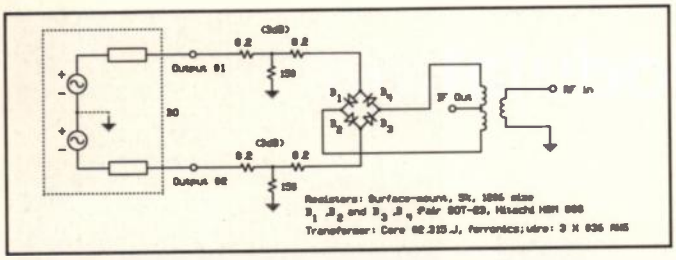

Figure 8. BO in the double balanced mixer circuit.

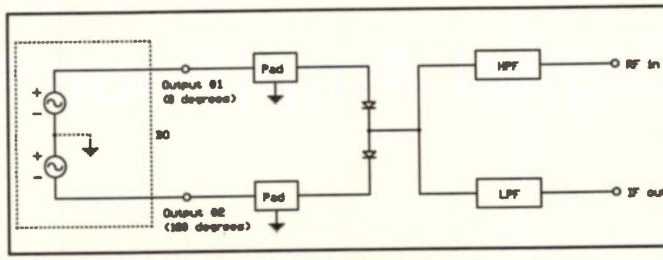

Figure 9. BO in a single balanced mixer circuit.

Conclusion

A balanced RF oscillator, with two out-of-phase outputs, possessing excellent amplitude and phase balance and isolation characteristics has been presented. The oscillator has 3 dB or more power advantage over single-ended oscillators and yet has substantially lower radiation and leakage levels. The power advantage makes room for the use of resistive attenuators, which further enhances the impedance and isolation characteristics of the BO. In addition, a BO has up to 3 dB better phase noise when compared with a single-ended oscillator. As with other oscillators, a BO frequency can be stabilized with elements such as crystals, SAW, dielectric resonators, etc.

A BO uses two active elements, as compared to one element used in single-ended oscillators. On the other hand, a single-ended oscillator usually needs a buffer amplifier, which makes the number of active elements equal. Furthermore, the cost of active elements is always dropping. So, by eliminating the need for balanced transformers, a BO has even a cost advantage over the single-ended one. For similar reasons, a BO can have a size advantage too.

The performance of the BO depends primarily on the matching of the active elements. For especially good performance, match elements, or even monolithic ICs, transistor arrays can be used. Equipped with all these characteristics, a BO is an excellent device for many applications. It is particularly useful from 1 GHz to 2 GHz, where the lumped and distributed circuit components overlap (and where neither of the two covers all the requirements) and where a good balanced transformer is very difficult to make. rf

About the Author

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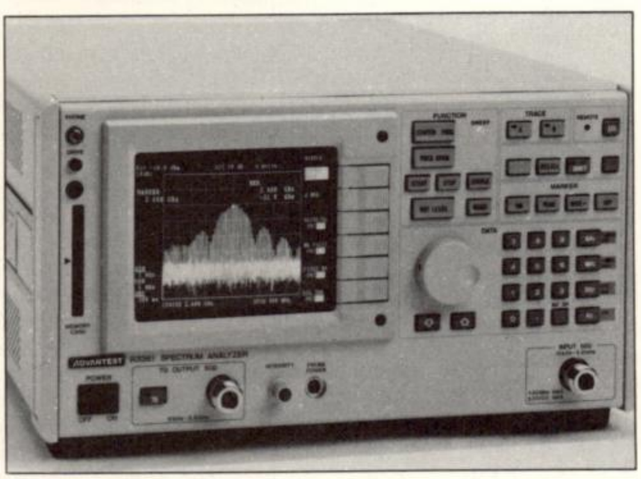

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Digital Amplitude Modulation

By Timothy P. Hulick, Ph.D.
Acrodyne Industries

Amplitude modulation is the oldest method of impressing information onto an electromagnetic carrier. Its beginnings date back to the early days of spark-gap transmissions. With spark came 100 percent amplitude modulation and the adoption of the Morse Code (later the International Morse Code), and still later, voice amplitude modulation, which has been shortened simply to AM.

AM usually refers to full-carrier amplitude modulation with a single set of inphase sidebands containing the information to be transmitted and received. Any modulation system that causes the instantaneous composite amplitude of the waveform to vary in accordance with the information transmitted is (or should be) termed AM. This includes single-side band suppressed-carrier emissions and vestigial-sideband television as well.

AM can be generated in many ways, but it always can be expressed by the familiar trigonometric identity:

$$(1 + \cos\omega_m t)\cos\omega_c t = \cos\omega_c t + (1/2)\cos(\omega_c - \omega_m)t + (1/2)\cos(\omega_c + \omega_m)t \quad (1)$$

where ω_m = modulation frequency,
ω_c = carrier frequency and
t = time.

This relationship commonly applies when the modulating waveform is a simple sine wave. More complex modulating waveforms may be expressed as a Fourier series of sine or cosine terms, but carrier and sideband terms retain the same form, and the modulation coefficient m modifies the amplitude of the $\cos\omega_m t$ term.

Any approach to develop the carrier and associated sidebands of equation 1 is fair game as a method to generate AM. The purpose of this article is to present a new method of generating pseudo-continuous amplitude modulation at any carrier frequency and at any modulation (depth) index between zero and one by using any class of amplifier (A, AB, B, C, D, H, S) as an RF source. First, however, it is necessary to review the operation of a common, yet often unfamiliar, RF component.

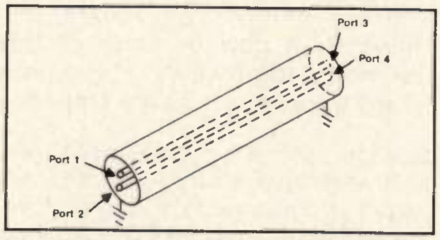

Figure 1. A quadrature hybrid is a 4-port device with two coupled lines inside a common outer conductor. The outer conductor is normally grounded.

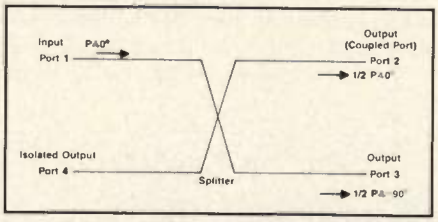

Figure 2. A schematic representation of the quadrature hybrid power splitter. If power is fed into port 1, it is split equally between ports 2 and 3 with the phase relationship shown. Ideally, no power leaves port 4.

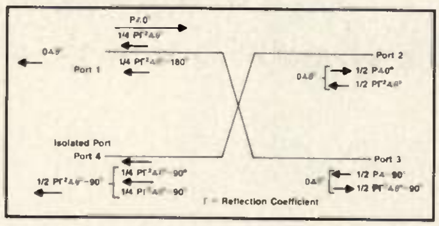

Figure 3. For the same type of mismatch at ports 2 and 3, all reflected power is transferred to port 4. Reflected powers to port 1 cancel each other because of phase relationships shown.

The Combiner/Splitter

The quadrature hybrid power combiner/splitter is well-known in RF circles. It is favored for microwave systems because of its small size at those frequencies and because it provides a practical way to sum the RF output power of many signal sources, produc-

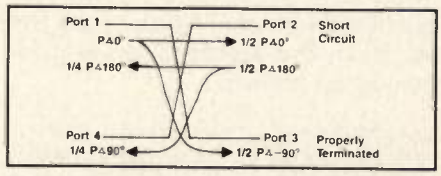

Figure 4(a). Output ports 2 and 3 are isolated from one another's mismatch. The signal path shown is when port 2 is terminated in a short.

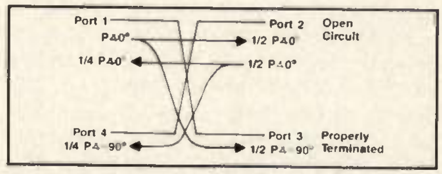

Figure 4(b). The paths when port 2 is terminated as open.

ing a much larger signal than that available from any single source. The device seldom is found in RF designs below about 50 MHz, because the size and cost become prohibitive. Its theory remains valid at all RF frequencies, however. The device may be referred to as a hybrid, a combiner, a splitter or a combination of these terms.

A quadrature hybrid combiner is a 4-port component that consists of two or more parallel conductors placed inside, but isolated from, a common outer conductor, such that the two lines share the same E and H fields. For this basic definition, disregard any restrictions on such factors as characteristic impedance of the coaxial arrangement or the location of terminations. Simply stated, two conductors that share a common field mutually induce current in one another according to laws of physics. The TV section of the *NAB Engineering Handbook* contains a vector analysis of the device.

The hybrid device exhibits several interesting and significant properties. Figure 1 shows coupled lines at the center of a common outer (grounded) conductor. Four ports, where appropriately sized connectors may be attached, are identified. The hybrid is shown schematically in Figure 2.

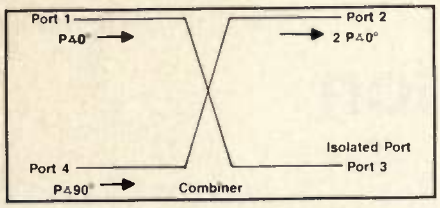

Figure 5. A schematic representation of the quadrature hybrid as a power combiner. The device is the same as the splitter, but is connected as shown.

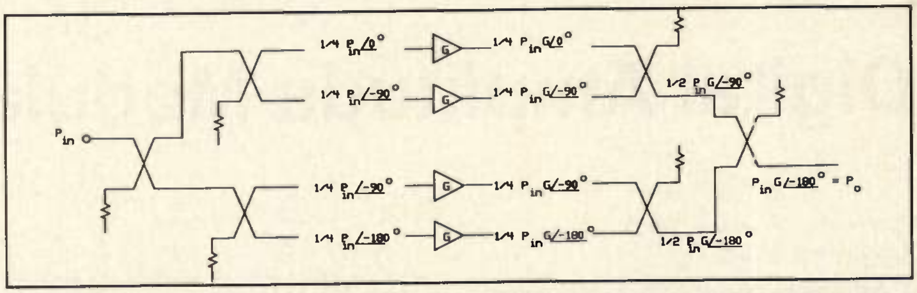

Figure 6. A configuration of three splitters and three combiners is connected to preserve phase. In the output signal,the four gain blocks (g) will appear as a single amplifier.

If the device, configured as a splitter, is of the correct dimensions for a given frequency, then the power of an input signal applied to port 1 is divided equally between ports 2 and 3. The signal at port 2 exhibits the same phase as that at port 1, excluding small propagation delays. At port 3, the phase is −90 degrees with respect to ports 1 and 2.

Whether the power is split equally between ports 2 and 3 depends upon the electrical length of the lines and the degree of coupling, which, in turn, are related to the shape and proximity of the lines. The characteristic impedance (Z_c) depends upon the cross-sectional geometry of the entire structure. If the lines and outer conductor are of circular cross section, the ratio of the inside diameter of the outer conductor to the outside diameter of one of the lines should be 4 for Z_c = 50 ohms at each of the four ports. If the lines are circular, and the outer conductor is square in cross section, then the ratio of the inside length of one of the sides of the outer conductor to the outside diameter of one of the lines should be 3.5 for Z_c = 50 ohms.

The degree of coupling depends upon the spacing between the lines, while the length of the enclosed line determines the frequency range over which the degree of coupling remains reasonably constant. The device maintains a degree of coupling to within a few tenths of a decibel more than an octave of bandwidth (f to 2f) and a nearly constant 90 degree phase shift to the quadrature port. Outside the octave bandwidth, coupling decreases in both directions, and the phase angle departs greatly from 90 degrees.

Terminations

If ports 2 and 3 are terminated in the proper characteristic impedance, no power is coupled to port 4. On the other hand, if the same magnitude and phase mismatch exist at both output ports 2 and 3, then they effectively become input ports for the reflected waves produced. As shown in Figure 3, the mismatch conditions previously described cause all reflected power to appear in port 4. Port 1, however, does not see the mismatch conditions, so a perfect termination is maintained at the expense of power lost to port 4.

If the mismatches at ports 2 and 3 are not alike, some power is reflected back to port 1, because amplitude and phase cancellation cannot occur at port 1. This presents a problem if loads connected to port 2 or 3 of a hybrid power splitter change in some way. The problem in the splitter configuration is less serious than it might appear, however, because power levels to the splitter generally are much lower than those to a combiner.

Port 4 is called the isolated (or reject) port, where a dummy load is connected to absorb reflected power. It is sized according to the expected worst-case reflected power. If the loads connected to ports 2 and 3 were always perfect, a dummy load at 4 would not be needed, and the port could be left unterminated.

That the two output ports ideally do not see each other is a significant property of the hybrid. A mismatch may occur at one output port, yet the other sees no reflected power. This is true (see Figure 4) whether a port is short or open-circuited. It is possible because, for a signal returning into port 2 (or 3), the opposite output port 3 (or 2) becomes the new isolated port for the reflected wave. With all other ports properly terminated, no power goes to the new isolated port. It appears that the output ports are isolated from each other, but the real degree of isolation is within the range of 20 dB to 30 dB.

Signal Combining

Figure 5 illustrates the hybrid as a power combiner, corresponding to the splitting configuration of Figure 2. Because the hybrid is a reciprocal device, the analysis is the reverse of the splitter. Two equal-amplitude signals with a split in phase of 90 degrees are applied to ports 1 and 4. They combine into port 2, while port 3 becomes the isolated port. It retains all the properties of the splitter, likewise affording input ports isolated with respect to one another.

If a hybrid is constructed to be a 3 dB splitter or combiner, power is equally split in amplitude to two ports or combined completely from two equal-amplitude ports. If power levels at the two combiner input ports are unequal, some power will be lost in the isolation-port load. Because vector voltages are combining to produce power, the output P_O is related to the input power levels P_{IN1} and P_{IN2}, according to the following relationship:

$$P_O = ((P_{IN1}/2)^{1/2} + (P_{IN2}/2)^{1/2})^2 \qquad (2)$$

The power sent to the isolation port becomes:

$$P_{ISO} = ((P_{IN1}/2)^{1/2} - (P_{IN2}/2)^{1/2})^2 \qquad (3)$$

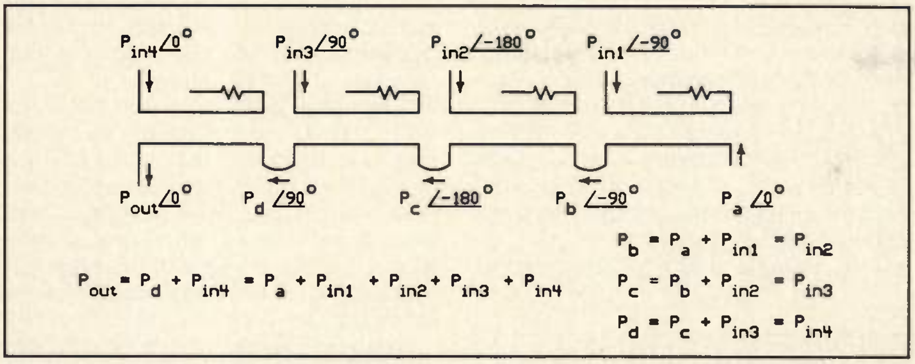

Figure 7. Four power-combining hybrids are connected with the output of each, feeding one input port of the unit to its immediate left. Input power at each input port doubles from the previous one, moving from the right, so that they are summed at P_{OUT}.

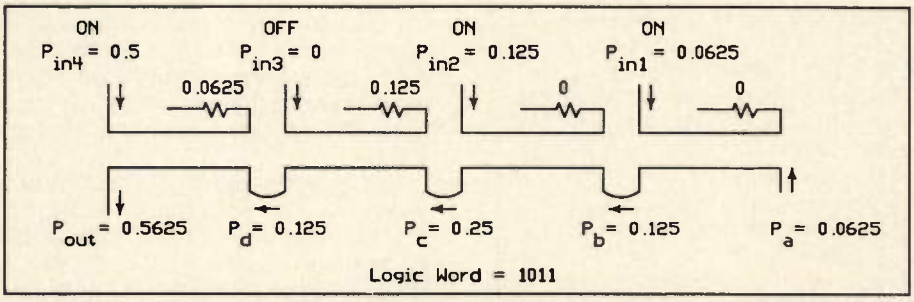

Figure 8. Combining hybrids with P_{IN2} in the OFF condition. Power levels shown represent a relationship to a total power of one unit if all inputs were ON.

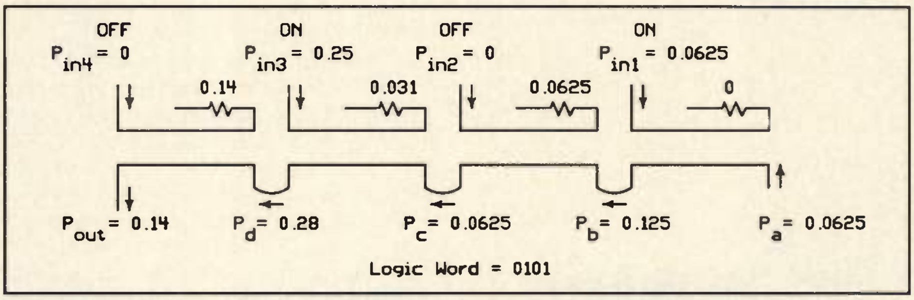

Figure 9. The system illustrated with P_{IN3} in the OFF condition.

If $P_{IN1} = P_{IN2}$, then equation 2 reduces to the sum of the input powers, while equation 3 goes to zero. If either P_{IN1} or P_{IN2} is zero, half the power of the remaining active input goes to the output port, while the other half appears at the isolation-port dummy load. The input ports remain isolated from one another.

Multiple quadrature hybrid sections may be connected in various ways to achieve a desired purpose. Figure 6 illustrates an interconnection to split drive power among four output ports to drive four separate amplifiers. The outputs of the amplifiers are applied to a combiner configuration, summing the signals back to a single output port.

Cascades of Digital Gates

Combining properties of the basic quadrature hybrid and relationships (equations 2 and 3) produces the configuration shown in Figure 7. The symbol for this application of the hybrid is changed to more closely reflect the actual construction of the device and is helpful in preventing crossed lines interconnecting the cascaded combiners. In all hybrids shown, the dummy or reject load is connected to the isolated port with respect to the two output ports.

Power input doubles in moving from right to left, so that $P_{IN2} = 2P_{IN1}$, and $P_{IN3} = 2P_{IN2}$, and so on.

System power levels are selected so that they sum to one, but the most important relationship is that the input power levels are weighted relative to the placement of digits of a binary number made of ones and zeros. Moving from the right in a binary word, each succeeding character has twice the numerical weight of the number to the immediate right. In Figure 7, all inputs are on. That is, if input port activity is controlled by a one or a zero digit of a binary word, they all would be the same — one or zero. (For the sake of simplicity, logic one conventionally means on, and logic zero corresponds to input power off.) P_a always exists and is not considered to be controlled by a binary digit, but represents the total power coming from all previous stages, if they exist.

The result is an amplitude modulator that can be controlled by the binary representation of the instantaneous value of the amplitude of an arbitrary waveform. Obviously, a mixture of ones and zeros results from any sample taken. For a modulator of this type, it is necessary for the power output of the combiner to produce the proper RF power level representing the modulation level sampled. Otherwise, modulation will be non-linear, and distortion will occur.

Figures 8 to 12 show numerical examples of the resultant summed power with one or more inputs turned off by a logical zero, assuming the same configuration as shown in Figure 7. For simplicity, assume the summed power to be one unit of power (1 W). Four input ports allow a 4-bit word of 16 possible states.

The lowest power input becomes one divided by 16, or 0.0625 W. Equations 2 and 3 allow intermediate power levels to be found. Figure 8 indicates that a logical word of 1101 produces an output level of 0.765 power units, while the input sum P_{IN} is 0.875. In Figure 9, a control word of 1011 produces the output 0.5625 from an input total of 0.75. Figure 10 shows an output power of 0.25, with an input sum of 0.5.

It is worth noting (and essential to the linearity of the modulator) that the output power is numerically equal to the square of the input power (0.8752 = 0.765, 0.752 = 0.5625, 0.52 = 0.25). If carried through for all 16 cases of the 4-bit word, the relationship holds. At first, it appears the square law relationship would render the combiner useless as a linear amplitude modulator. However, the digital voltage representations of an arbitrary waveform are just that — voltages controlling powers that automatically square the voltages into powers so that the squaring is canceled. In effect, the digital voltage word is squared into a power word that is perfectly linearly proportional to the power at the summed port P_{OUT}.

The linearity of the modulator is independent of the type of power source, as long as available power to each port remains precisely double that of the next lower power-input port. The proper amount of waste power automatically finds its way to a reject load, and not to P_{OUT}, to maintain linearity.

It also may seem, at first glance, that

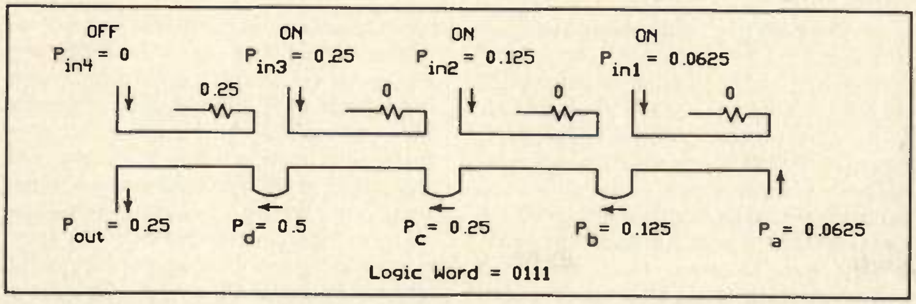

Figure 10. The system illustrated with P_{IN4} in the OFF condition.

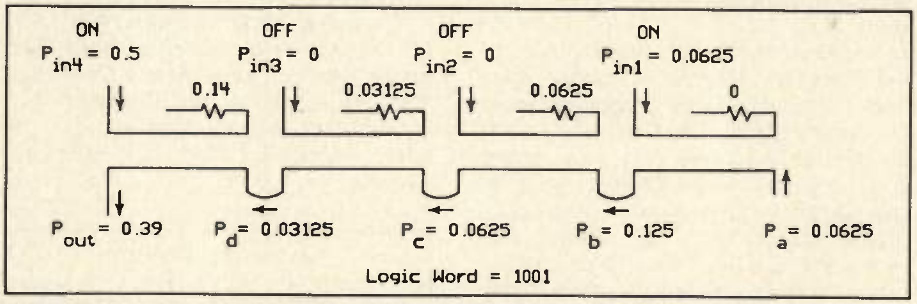

Figure 11. The system with P_{IN2} and P_{IN3} in the OFF condition.

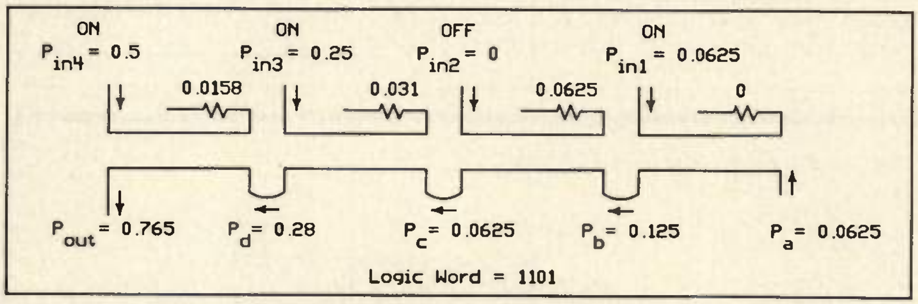

Figure 12. The system with P_{IN2} and P_{IN4} in the OFF condition.

the modulator is terribly inefficient, because power must be dumped into reject loads for the system to work. The combiner/modulator is theoretically 100 percent efficient with all inputs on. Efficiency decreases as some inputs change to the off condition, but so does total consumption.

A Technological Concept

Digital technology is making significant inroads into the broadcast industry. As that technology moves more toward signal processing and transmission, the approach to digital modulation discussed in this article allows for analog simulation of a digitized signal in an amplitude-modulated broadcast system. Reception is possible with ordinary radio or TV receivers.

Digital AM transmitters for medium-wave applications exist, but the technology that makes them possible uses a different approach than that presented here. Power combiners in those transmitters do not offer port-to-port isolation. That is, no reject loads are associated with the power combiner, which acts as a transformer with a single secondary and many primaries. Also, an isolated combiner is not necessary at medium wave, where switch-mode RF amplifiers form the modulator/RF source. Such a source impedance is either near zero when gated on, or approaching infinity when gated off. Switch-mode amplifiers are not yet possible much above the medium-wave frequencies, suggesting that another method is required.

The modulator described here, because of the port-to-port isolation, maintains an impedance of R_L at all ports whether or not adjacent amplifiers are on. As a result, it is useful at all RF frequencies with any class of amplifier. Its speed is limited only by the ability to gate an amplifier in consonance with the analog-to-digital converter sampling rate. No doubt, logic glitches may result as amplifiers are turned on and off, but a careful design should overcome this problem.

The modulator described would not work for vestigial-sideband transmission in its current configuration. The technique for partial lower-sideband cancellation is not developed at this time, nor is it known to exist, but it deserves future consideration. Even if it does not exist, a high-level vestigial filter could be constructed to pass the appropriate spectral components of the double-sideband signal for NTSC TV transmission.

Because of the absolute linearity of the modulator, it is expected that usual non-linearities of TV RF amplifiers would be non-existent. Non-linearities, such as differential phase and gain, group delay and low- and high-frequency response seem to disappear. Analog-to-digital circuitry with anti-aliasing and replicating spectra filters to counteract the sampling process is deliberately omitted from this discussion because those circuits and processes are well-known. They would apply equally to the output bandpass filter to suppress the radiation of spurious signals.

The application of this modulator would be a reverse trend to high-level modulation in TV transmission. It is more than a modulator, however. It is a transmitter in which the modulator and RF power amplifier sections are inherently one. Through logic and numerical analysis, the advantages of digital technology appear to be applicable to the transmitting system as well. Perhaps in the future, an operating model based on these concepts will prove the theory.

The author received Patent 4,804,931 for the modulator described here. This article was published in the December 1987 issue of Broadcast Engineering. *It is reprinted with permission from Intertec Publishing Corp.*

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A New Broadband Antenna

By Roger Southwick
EMC Consulting

This article introduces a new antenna called the bowtop antenna. It is an active receiving antenna that is broadband and is linearly polarized. The antenna has various advantages, which make it suitable for EMC related test applications such as the FCC Part 15/J.

The bowtop antenna features a 30 MHz to 700 MHz bandwidth with an antenna factor of 3 dB ±5 dB. Its dimensions include a height of 19 cm and a diameter of 14 cm. The bowtop antenna with mounting boom is shown in Figure 1. A calibration curve is shown in Figure 2. Note that the bandwidth can be extended to 1000 MHz.

The bowtop antenna element is one element of a bowtie antenna with a circular top load and an attached ground plane. The antenna structure consists of a triangular-shaped element made of copper foil; it is attached to a plastic backing and mounted inside a plastic tube. The circular top load is also made of copper foil attached to the underside of the plastic top of the tube. The element and the top load are connected with copper tape and soldered together at the base of the triangle. The bottom of the tube is a circular aluminum plate that acts as a ground plane. The outer plastic tube provides protection for the actual element, making it extremely difficult to damage. In the side of the plastic tube is a rectangular opening with a bracket on either side which is used to mount the amplifier.

The bowtop element design is based on the Bifin antenna design. One limitation of a bowtie antenna is that at lower frequencies, its impedance rapidly becomes capacitive, thereby causing a dramatic reduction in efficiency. The bowtop antenna solves this problem by adding a top load. This combination extends the frequency range so that an element length of only 15 cm is usable to 30 MHz.

The effect of the top load is best understood in the monopole case. As with all other linear antennas, the bowtie antenna can be used in either a monopole or a dipole configuration. Consider a monopole configuration with a single 15 cm element above a ground plane. In such a case, the bowtop antenna acts like a bowtie antenna at higher frequencies, like a top loaded antenna down towards 50 MHz, and like a disk antenna at still lower frequencies. That is, the bowtop antenna combines characteristics of these three antenna types into a single physical configuration. The result is a physically small antenna, with bandwidth characteristics far greater than any other single configuration.

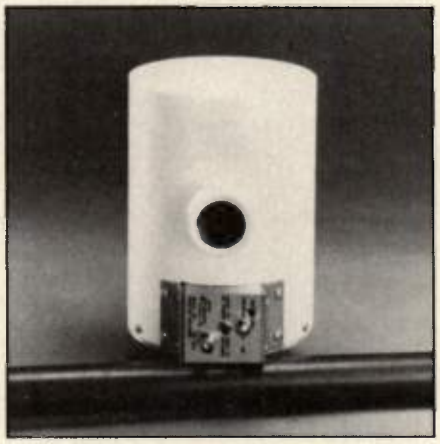

Figure 1. The bowtop antenna with mounting boom.

Unlike the classic dipole antenna, which includes a balun to combine the inputs of the two elements, the bowtop antenna is unbalanced and therefore no balun is required. The output of the single element is connected directly through an RC (resistive/capacitive) network to the low-noise amplifier. As the single-element antenna depends on the top load at its lower frequency, the metal base plate is included in the antenna structure to provide the necessary capacitance to the top load. Although this arrangement is somewhat less efficient than that of a dipole, it eliminates the need for a balun. Further, this configuration can be used in free space in the fashion of a dipole antenna because the base plate allows it to be independent of earth ground.

The Amplifier

The active amplifier and power source are contained in a small aluminum case that fits into the opening in the antenna element. At the apex of the triangular antenna structure is a jack, which is the element output, and below it is a second jack, which connects the ground plane base plate. The amplifier case has a pair of matching jacks, one to ground and the other to the amplifier input. When the case is inserted into the opening in the element, the jacks are connected, the antenna element is connected at the amplifier input, and the aluminum case and base plate ground are connected. Four screws on the bracket secure the amplifier case to the element. A cutaway view of the bowtop antenna is shown in Figure 3.

The case contains an amplifier, two 9 volt batteries, and a voltage monitor. The amplifier gain is shaped by an RC network at its input, so that the element mismatch is compensated for by the amplifier gain to provide a near constant antenna factor across the antenna range. The amplifier has two RF transistors, and has a noise figure of 4 dB. The voltage monitor cuts off the amplifier and turns on a warning light when the battery voltage drops below a preset level.

The bowtop antenna is extremely simple to operate since no adjustments or calibration procedures are required. On the front of the case is an ON-OFF switch, a BNC output connector and a DC power jack. The bowtop antenna may be operated from either the internal batteries or an external DC power supply. The batteries provide about ten hours of continuous operation and are rechargeable through the DC power jack.

Operation

As mentioned above, the bowtop antenna can be used as either a dipole or a monopole antenna. To facilitate dipole operation, a boom can be screwed into a mounting socket on the element tube. (An 18 in. boom is provided.) With the boom connected, the bowtop can be mounted on a tripod or antenna scanner to measure either vertically or horizontally polarized E-fields.

When the bowtop is used as a monopole antenna, it should be placed

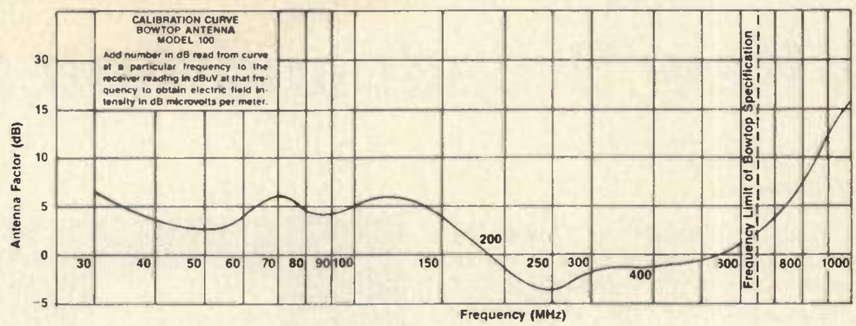

Figure 2.

Figure 2. The bowtop calibration curve.

on the ground plane with the base plate down. Because the bowtop antenna pattern is not omni-directional, the amplifier case must always be aligned opposite the source being measured.

Applications

Wide bandwidth antennas such as the bowtop are desirable for measurement applications, as it is not necessary to change antennas or switch bands. Such antennas are more adaptable to automated measurement systems in that a wider bandwidth can be automatically scanned. This avoids the problem of having to design the automated system around the antenna bandwidth. Further, it is very important for an antenna to have a nearly constant antenna factor when used in conjunction with an automated measurement system. When this is not the case, the field strength data is greatly distorted by the unevenness of the antenna factor. This situation will result in a similar unevenness of the sensitivity of the system in terms of field strength. However, the nearly constant antenna factor of the bowtop antenna minimizes this problem.

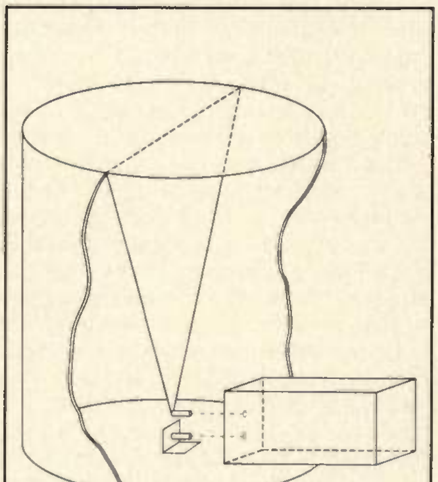

Figure 3. Cutaway view of the bowtop antenna.

Physical size is also an important consideration. Large antennas are not only difficult to handle and transport, but they also have a greater tendency to couple to nearby objects. In the case of the FCC Part 15/J scanning requirements, the larger antenna will couple to the ground plane in horizontal polarization and may limit the scan travel if the lower antenna element hits the ground plane or the chamber ceiling. Because of the bowtop's small physical size, coupling will be insignificant and it will meet the FCC Part 15/J scanning requirements. The bowtop is so small that it is ideal as a portable hand-held probe. This application is very useful in ISM work, or for situations in which measurements must be made on site or in a restricted space.

Surveillance is another application and, again, because of the bowtop's small size, it would be suitable for mounting either on the roof of a vehicle or on the surface of an aircraft. The wide bandwidth allows the bowtop to replace several conventional antennas in this application.

Further information on this antenna can be obtained from E. T A. Engineers, 3325 N. Forgeus Ave., Tucson, AZ 85716. Tel: (602) 323-8174.

About the Author

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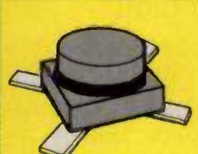
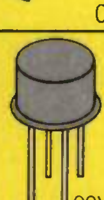
 CGY 40	Micro-X	17.5	10	2.9	1.4	32.5	20	100 MHz–1.8(4) GHz
 CGY 31/21	TO-12	19	19	3.8	1.5	34.5	30	40 MHz–1.8(3) GHz

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INFO/CARD 45

Design of Wideband Quadrature Couplers for UHF/VHF: Part II

By Chen Y. Ho,
M/A-COM Active Assemblies Div.
and Ge-Lih Chen, Chung Shan Institute of Science and Technology

This article presents the same design approach as the wideband quadrature coupler discussed in Part I (November 1989 RF Design*). It uses a lowpass filter of proper phase characteristics to replace the transmission lines which interconnect the narrowband quadrature coupler. The results are particularly useful for the lower frequency spectrum where transmission lines become excessive in length. Experimental results show good agreement with the theoretical computations.*

At the lower end of the VHF band, the physical length of the interconnecting transmission lines used in the wideband quadrature coupler becomes excessively long and in many cases they are impractical to realize. Since the main function of the transmission lines is to provide a proper phase relationship between the narrowband couplers, it can be replaced by lumped element lowpass filters with proper phase characteristics. Any circuit which has minimum insertion loss and mismatch and possesses proper phase characteristics over the frequency band of interest can be used to replace the transmission lines. Some degradation in performance can be expected due to imperfect match and the non-ideal linear phase characteristics of the filter. An advantage is that the size of the coupler can be considerably reduced.

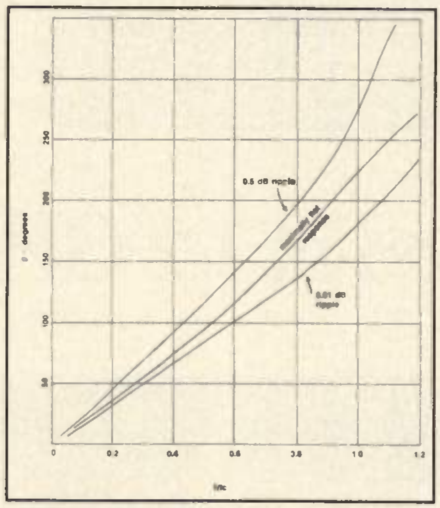

Figure 1. Phase shift characteristics with maximally flat or Chebyshev attenuation responses and n=5.

Phase Characteristics of a Lowpass Filter

The phase characteristics of a lowpass filter prototype (n = 5) are shown in Figure 1. It can be seen from the figure that the phase characteristic of a low ripple Chebyshev response filter design is almost the same as that of a maximally flat response design for $f \leqslant 0.6f_c$, where f_c is the cut-off frequency of the filter. It is known that filters with maximally flat responses have poorer

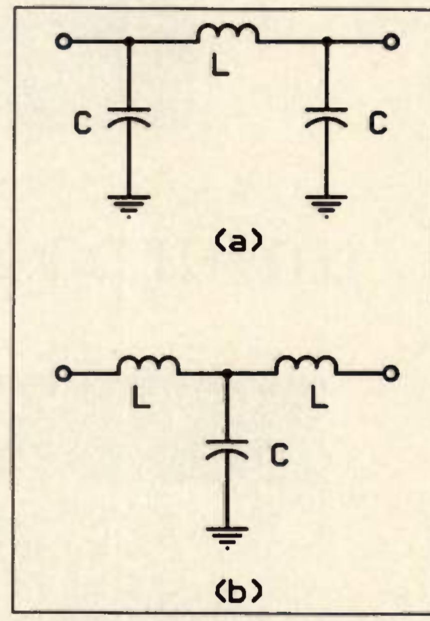

Figure 2. Lowpass filter configurations.

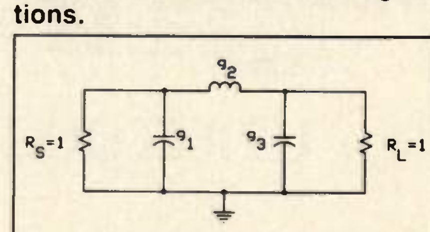

Figure 3. Lowpass filter prototype.

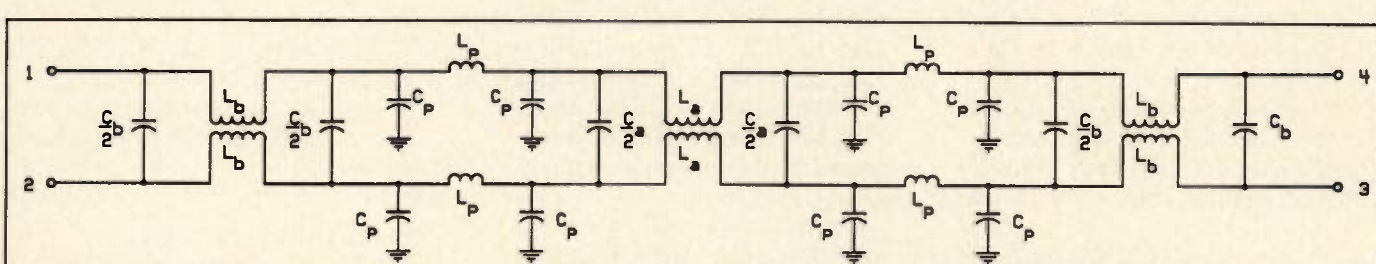

Figure 4. The wideband quadrature coupler with lowpass filters.

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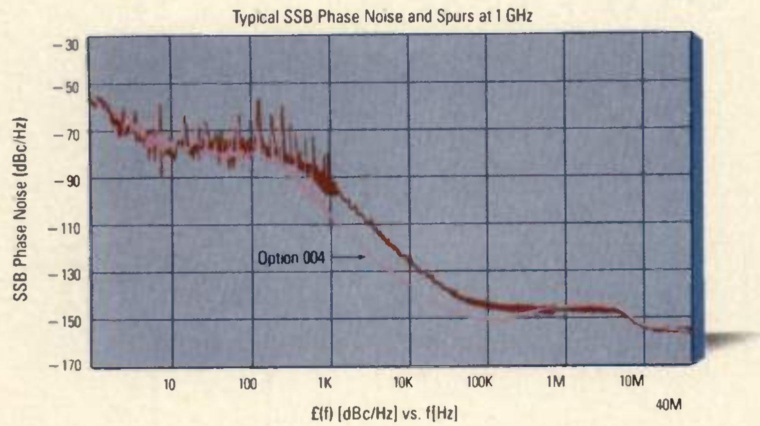

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return loss (0, f_c) for the same amount of ripple. Only filters with Chebyshev frequency response will be considered.

The total phase shift of an nth order filter is equal to $n\pi/2$ for frequencies from DC to infinite. However, most of the phase shift occurs from DC to f_c. For a filter with a Chebyshev frequency response, n should be an odd number in order to avoid an unequal termination for source and load. For n = 1, the amount of phase shift may not be sufficient to achieve the phase shift required to replace the transmission lines. For $n \geq 5$, the number of elements is large and it becomes difficult to build two or more lowpass filters of similar phase characteristics.

With the above consideration in mind, only filter designs with low ripple Chebyshev response and n = 3 are used for the replacement of the transmission lines. The passband ripple of a filter is directly related to the return loss in the same band.

The analysis for the filters in Figure 2 is similar. Hence, only the circuit in Figure 2(a) will be discussed. For the lumped element lowpass filter prototype shown in Figure 3, where $g_1 = g_3$ for symmetry, θ can be expressed as:

$$\theta = \tan^{-1}\left[\frac{g_2\left(\frac{f}{f_c}\right)}{2\left(1 - g_1 g_2\left(\frac{f}{f_c}\right)^2\right)} + \frac{g_1\left(\frac{f}{f_c}\right)\left(2 - g_1 g_2\left(\frac{f}{f_c}\right)^2\right)}{2\left(1 - g_1 g_2\left(\frac{f}{f_c}\right)^2\right)}\right] \quad (1)$$

where f_c is the cut-off frequency of the filter and g_1 and g_2 are the element values of the lowpass prototype. The required length of the transmission lines for the wideband quadrature coupler is θ_o at f_o. At $f = f_o$, equation 1 becomes:

$$\theta_o = \tan^{-1}\left[\frac{g_2\left(\frac{f_o}{f_c}\right)}{2\left(1 - g_1 g_2\left(\frac{f_o}{f_c}\right)^2\right)} + \frac{g_1\left(\frac{f_o}{f_c}\right)\left(2 - g_1 g_2\left(\frac{f_o}{f_c}\right)^2\right)}{2\left(1 - g_1 g_2\left(\frac{f_o}{f_c}\right)^2\right)}\right] \quad (2)$$

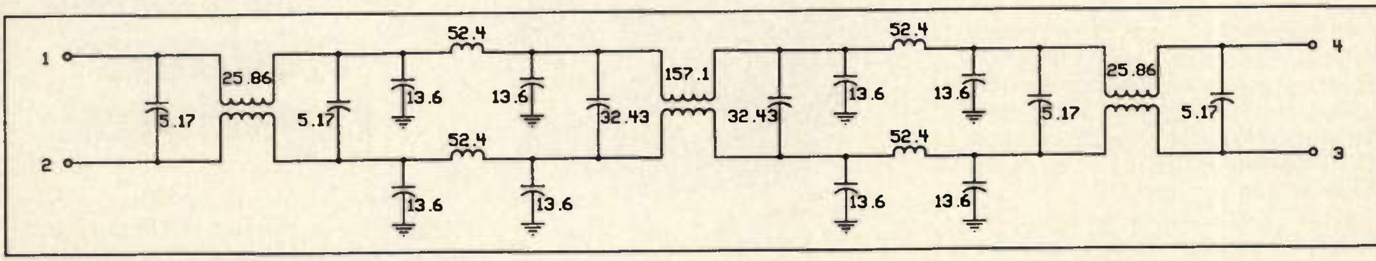

Figure 5. The coupler circuit.

If g_1 and g_2 are known for a particular design, f_c can be computed by solving equation 2. L_p and C_p are related to f_c by:

$$C_p = \frac{g_1}{2\pi f_c Z_o} \quad 3(a)$$

$$C_p = \frac{g_1}{Z_o(2\pi f_c)} = 13.6 \text{ pF} \quad 3(b)$$

By replacing the transmission lines with the lowpass filters, the configuration of the wideband quadrature coupler shown in Figure 4 is obtained.

Experimental Results

The same design examples used in Part I are used to demonstrate the design approach described here. For a wideband quadrature ccupler centered at 80 MHz, L = 99.47 nH, C/2 = 19.89 pF, a = 1.63, b = 0.26 and θ = 35 degrees. The lowpass filter chosen has 0.01 dB ripple or 26 dB return loss for $f \leqslant f_c$. This results in element values of $g_1 = g_3 = 0.6291$ and $g_2 = 0.9702$.

For θ_o = 35 degrees at 80 MHz, using equation 2, f_c is 147.34 MHz. The element values for L_p and C_p can be computed from equation 3(a) and 3(b):

$$L_p = \frac{g_2}{2\pi f_c}(Z_o) \quad 4(a)$$

$$L_p = \frac{g_2 Z_o}{2\pi f_c} = 52.4 \text{ nH} \quad 4(b)$$

A wideband coupler has been fabricated and tested. The circuit is shown in Figure 5. The coupler is fabricated using the same techniques as those used in Part I, except that the transmission lines

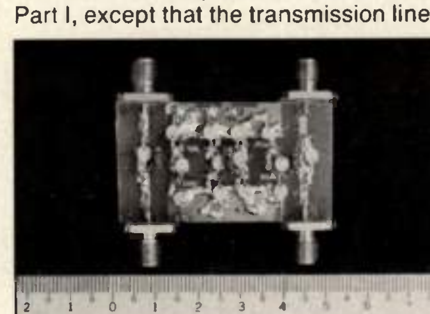

Figure 6. The coupler prototype.

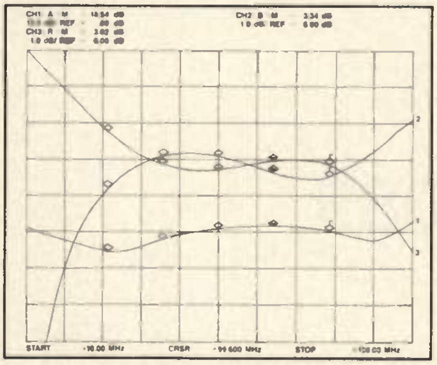

Figure 7. Measured frequency response.

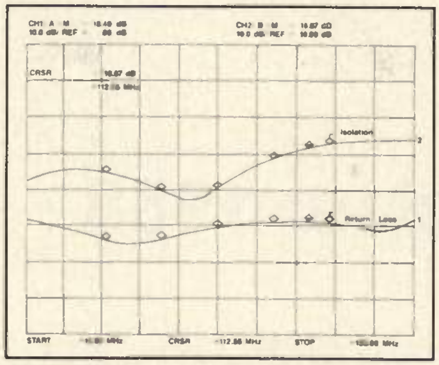

Figure 8. Return loss and isolation. of the coupler.

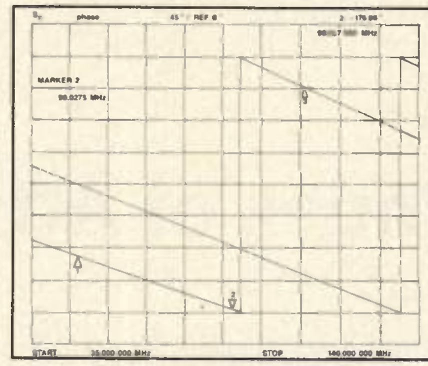

Figure 9. Quadrature phase relationships.

are replaced by lumped element lowpass filters. The capacitors used in the lowpass filter are ATC chip capacitors and variable capacitors from Johanson. The inductors are air-core winding, 5 turns, 0.250 in. I.D. The photo of the coupler is shown in Figure 6. The measured response (after some adjustments) of the wideband quadrature coupler is shown in Figure 7 (amplitude ripple response). Figure 8 illustrates the return loss and isolation, and Figure 9 shows the quadrature phase relationships. The measured ripple is about 0.5 dB from 47 MHz to 130 MHz or bandwidth ratio of 2.75. The return loss is better than 18 dB and isolation is better than 16 dB. ■

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About the Authors

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Amplifier Design

Oscillator Design

Filter Design

Transmission Lines/Couplers

Frequency Synthesis

Test and Measurement

Errata, June 1989, p. 13.
Increased Dynamic Range Measurements Using a Network Analyzer, Chris Day, May 1989, p. 44.
Principles of Digital Storage Oscilloscope Operation, Gene Andrews, May 1989, p. 52.
VXIbus: Benefits for RF Applications, Malcolm Levy, September 1989, p. 32.

Mixers, Multipliers and Dividers

Product Detector Performance, Fred Brown, March 1988, p. 47.
MMIC Active Multipliers, Jerry Hinshaw, June 1988, p. 64.
Unequal Power Division With a Lumped Element Divider, David Burgess, June 1988, p. 69.
A Mixer Spurious Plotting Program, Richard Bain, May 1989, p. 32.
Errata, July 1989, p. 13.
Frequency Division With Varactor Diodes, William J. Hoffert, October 1989, p. 35.
A Spurious Response Program for Wideband Mixer Circuits, William E. Sabin, October 1989, p. 41.

Modulation/Demodulation

A Loran-C Front End Using the LM1211, Ralph Burhans, August 1988, p. 23.
An Improved Double Sideband Modulator, George J. Misic, December 1988, p. 62.
Carrier Detection Utilizing FM Click Characterization, Gerald L. Somer, June 1989, p. 80.
Complex Signals: Part 1, Noel Boutin, December 1989, p. 27.
Digital Amplitude Modulation, Tim Hulick, December 1989, p. 39.

Antennas and Propagation

Wide-Band Test Antennas, Theodore Rappaport, April 1988, p. 37.
Plate Antenna Couplers, T. Koryu Ishii, September 1988, p. 24.
Multipath Interference in FM Data Transmission, Steven E. Turner, December 1988, p. 26.
Simplifying Propagation Loss Calculations, Mel Bebee, 1988 RF Design Directory, p. 65.
Factory Radio Communications, Theodore S. Rappaport, January 1989, p. 67.
RF Propagation in Buildings, T. Koryu Ishii, July 1989, p. 45.
Errata, October 1989, p. 12.
Active DF Antennas, Henry W. Anderson, R. Stephen Smith, November 1989, p. 62.

Circuit Analysis

Distortion in Nonlinear Circuits, Andrzej B. Przedpelski, February 1988, p. 61.
Errata, April 1988, p. 10.
Feedforward Compensation for Improved DC Performance, Stan Goldman, February 1988, p. 85.
Errata, May 1988, p. 8.
System Noise Analysis Using Spreadsheets, John H. Bordelon, David R. Hertling, November 1988, p. 56.
Design of Constant Phase Difference Networks, Thomas A. Keely, April 1989, p. 32.
BASIC Program for 90-Degree Allpass Networks, Allan G. Lloyd, April 1989, p. 47.
S-Parameters in Spice, Thomas B. Mills, June 1989, p. 45.
Software-Assisted Characterization of Large-Signal Device Models, Jeremy M. Bunting, September 1989, p. 27.
Calculating S-Parameters From Nodal Analysis, Bert K. Erickson, September 1989, p. 44.
A Nodal Network Analysis Program, November 1989, p. 82.
Principles of Transistor Interactive Mapping, Charles R. Ruelke, 1990 RF Design Directory, Fall 1989, p. 81.

High Speed Analog and Digital Circuits

High Speed A/D and D/A Conversion, Gary A. Breed, January 1988, p. 39.
Monolithic DAC for RF Signal Processing, John Sylvan, January 1988, p. 44.
A Review of Op Amp Specifications, Mark Gomez, March 1988, p. 49.
A Programmable Pulse Driver Formatter, Karl Zabel, April 1988, p. 26.
Errata, June 1988, p. 13.
Avoiding Ground Problems in High-Speed Circuits, Jeff Barrow, July 1989, p. 32.
High Frequency Design With Analog Master Chips, John Shier, April 1988, p. 48.
Noise Power Ratio Characterizes A/D Performance for Communications, Michael Johnson, October 1989, p. 44.
High Frequency Op Amp Makes Precise Detector and AGC, Brian Mathews, David Riemer, November 1989, p. 76.

Phase Shifters/Matching Networks

Lumped Element Phase Shifting and Matching Networks, Andre Boulouard, July 1988, p. 58.
Errata, October 1988, p. 13.
A Voltage-Controlled Phase Shifter, Thomas R. Mathews, October 1988, p. 38.
Low Impedance Double-Tuned Matching Circuit, Andrzej B. Przedpelski, 1988 RF Design Directory, p. 61.
Tapped Double-Tuned Matching Circuit (Bandpass), Andrzej B. Przedpelski, 1988 RF Design Directory, p. 62.
Response of Low Impedance Double-Tuned Matching Circuit, Andrzej B. Przedpelski, 1988 RF Design Directory, p. 63.
Dual Trap Frequency Response, Andrzej B. Przedpelski, 1988 RF Design Directory, p. 64.
CAD for Lumped Element Matching Circuits, Stanley Novak, February 1989, p. 102.
Design of Line Matching Networks, Peter Martin, February 1989, p. 99.
Errata, May 1989, p. 13.
Phase Relationships for Maximum Power Transfer, Robert A. Witte, June 1989, p. 70.

Miscellaneous Articles

A Proximity Warning System for Light Aircraft, H. Paul Shuch, May 1988, p. 43.
Quartz Devices Conference & Exposition: The Technical Program, August 1988, p. 38.
RF Components at Low Temperatures, Otward Mueller, January 1989, p. 29.
A Microwave Oven to Amateur TV Transmitter Conversion, David Pacholok, March 1989, p. 48.
A Simple Low-Cost RF Switch, Tad Harris, July 1989, p. 53.
A Simple 49 MHz Transmitter, Gary Carroll, September 1989, p. 38.
Modular RF Systems Shorten Design-to-Product Time, Alan Victor, October 1989, p. 31.
Perspectives on Ceramic Chip Capacitors. Part I: Device Properties, Mark W. Ingalls, November 1989, p. 49.
The RF Technology Expo 88 Technical Program, January 1988, p. 47.
The RF Expo East 88 Technical Program, October 1988, p. 54.
The RF Technology Expo 89 Technical Program, January 1989, p. 59.
RF Expo East Technical Program Offers Variety and Depth, October 1989, p. 50.
New Products Featured at RF Technology Expo 88, February 1988, p. 68.
New Products Featured at RF Expo East, November 1988, p. 76.
New Products Featured at RF Technology Expo 89, February 1989, p. 90.
New Products Featured at RF Expo East, November 1989, p. 70.
Profile of an RF Engineer, Gary A. Breed, July 1988, p. 43.
Looking Ahead: The Next Ten Years, J.A. Arden, Thomas B. Mills, Andrzej B. Przedpelski, December 1988, p. 49.
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Book Reviews

Single-Sideband Systems and Circuits, William E. Sabin, Edgar O. Schoenike, McGraw Hill; Reviewed by Gary A. Breed, July 1988, p. 61.
The ARRL Antenna Handbook, Gerald Hall, ARRL; Reviewed by Gary A. Breed, February 1989, p. 77.
Errata, May 1989, p. 13.
Notes on Microwave Circuits (3 vols.), Darko Kajfez, Kajfez Consulting; Reviewed by Gary A. Breed, February 1989, p. 77.
Digital Signal Processing Design, Andrew Bateman, Computer Science Press; Reviewed by Gary A. Breed, February 1989, p. 78.
Communication Receivers: Principles and Design, Ulrich L. Rohde, T.T.N. Bucher, McGraw Hill; Reviewed by Gary A. Breed, February 1989, p. 79.
Electronic Communications Handbook, Andrew F. Inglis, McGraw Hill; Reviewed by Gary A. Breed, February 1989, p. 79.

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HENRY FORD — DETROIT INDUSTRIALIST

☐ **A CHICKEN IN EVERY POT**
HENRY IV — KING OF FRANCE

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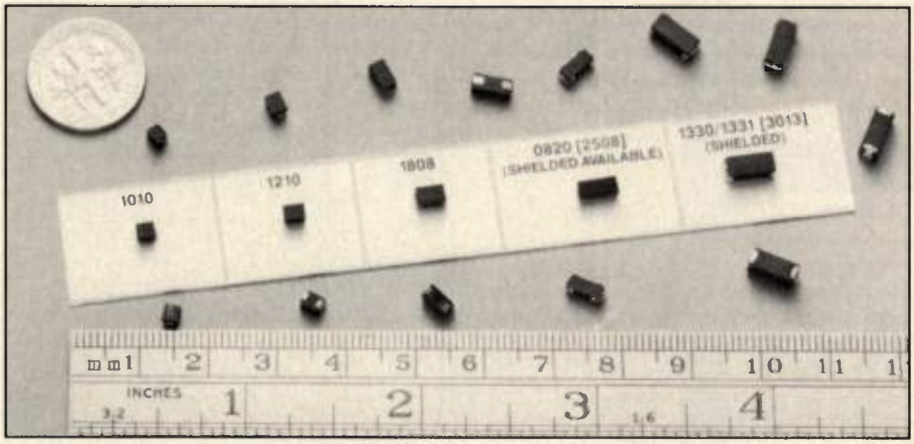
INFO/CARD 54

SMT Inductors From Delevan

The Series 1808, 1210 and 1010 inductors are available for inductances from 0.010 to 150 uH. They can be used to incorporate all the advantages of SMD technology into existing through-hole or new board designs.

The inductors feature J termination in a molded envelope to provide higher performance than conventional termination methods. Laser markings are used to identify the new series of inductors. They are priced at approximately 35 cents each in 1,000,000 piece quantity. **American Precision Industries, Delevan/SMD Division, E. Aurora, NY. INFO/CARD #230.**

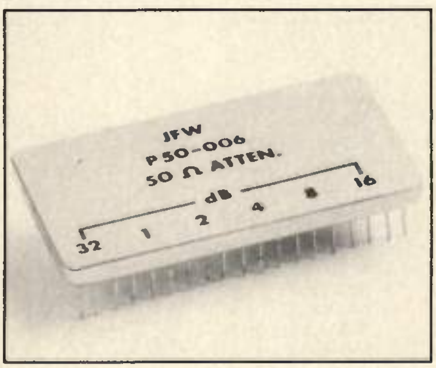


A Plug-In Hybrid Programmable Attenuator From JFW

Model P50-006 is a plug-in hybrid programmable attenuator with a 10 to 600 MHz frequency range. The attenuation range is 0 to 63 dB in 1 dB steps. VSWR is 1.5:1 maximum, insertion loss is 3.5 dB nominal and attenuation accuracy is ±0.3 dB or 2 percent, whichever is greater.

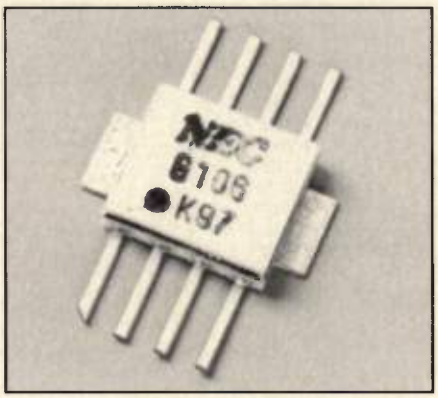
6-bit programming logic is TTL-compatible, with attenuation steps of 1, 2, 4, 8, 16 and 32 dB. The power requirement is +5 VDC at 250 mA.

Other specifications include a switching speed of 5 us, input power of +5 dBm and nominal impedance of 50 ohms. Temperature range is 0 degrees C to 85 degrees C, and the attenuator is 2 in. long and 1 in. wide. **JFW Industries, Inc., Indianapolis, IN. Please circle INFO/CARD #229.**



CEL Introduces an AGC Amplifier

The UPG106 wideband AGC amplifier from NEC is available in both chip (APG106P) and hermetic package



(UPG106B). Frequency range is 100 kHz to 2.5 GHz, typical gain is 20 dB and typical control range is 35 dB. Input and output impedance is matched to 50 ohms. The device has a reliable operating temperature range from −65 degrees C to +125 degrees C.

Applications include electro-optical and laser systems as well as local area networks, instrumentation and microwave communication systems. MIL-screened versions for satellite communications, EW and radar systems are available. Price ranges from $45 for the chip version to $68 for the hermetic package, in 100 piece quantity. **California Eastern Laboratories, Inc., Santa Clara, CA. INFO/CARD #228.**

New Conductive Adhesive Coating

AC84 is a one-component conductive adhesive coating designed for EMI/RFI shielding applications. It offers attenuation of 60 to 70 dB from 1 MHz to 1 GHz.

The coating has good adhesive qualities and can easily be applied to adhere tenaciously to most surfaces, including plastics, without the need for pretreatment. Application methods include air or airless type spraying equipment, brushing, dipping and silkscreening. Uniform continuous coatings are characterized by high electrical conductivity. Also, superior electromagnetic shielding effectiveness, which is resistant to weathering, abrasion, humidity and corrosion, is readily formed.

The coating is composed of an oxidation resistant and chemically resistant adhesive polymer. Temperature range is −80 degrees F to +300 degrees F. It can be used directly from the containers since no catalysts or accelerators need to be added. **Master Bond, Inc., Hackensack, NJ. INFO/CARD #227.**

Exhibitors

Acrian, Inc.
Advanced Electromagnetics, Inc.
Advantest America, Inc.
AEL Defense Corp.
Alan Industries
Alliance Technologies, Inc.
Alpha Industries
AML
AMP Incorporated
Amperex Electronic Corporation
Amplifier Research
Andersen Laboratories
Andrew Corporation
Anritsu
Applied Dielectrics
Applied Engineering Products
Avantek, Inc.
AVX Corporation
Ball Efratom Division
Balo Hermetics
Bird Electronic Corporation
Boonton Electronics Corp.
Burr-Brown
C.H.P.S. (Complete Hybrid Parts Supply)
California Eastern Labs
Cascade MicroTech, Inc.
Ceramic Devices, Inc.
Circuit Busters
Coaxial Dynamics, Inc.
Colorado Data Systems
Compac Development Corp.
Compact Software Inc.
Comstron Corporation
Connecting Devices Inc.
Cougar Components
Croven Crystals
Crystal Technology, Inc.
CTS Knights
Delta Microwave Inc.
Digital RF Solutions
Doty Scientific, Inc.
Dow Key Microwave Corporation
EDO Corporation/Barnes Engineering Division
EEsof, Inc.
Electro Rent Corporation
Electronic Decisions Inc.
EMC Technology, Inc.
ENI, Inc.
Erotec Engineering, Inc.
FEI Microwave
Ferretec Inc.
Fetty-Schoenduve, Inc.
Fotofabrication Corp.
Gamma-F Corp.
GigaBit Logic Inc.
Harris Microwave Semiconductor
Hewlett-Packard Company
Hewlett Packard Company— Microwave Semiconductor
Hopkins Engineering Co.
Huber & Suhner Inc.
Hughes Aircraft Co., Microelectronic Circuits
IFR Systems, Inc.
Innovative Frequency Control Products Inc.
Instrument Specialties
Instruments for Industry, Inc.
Intech Incorporated
Integrated Microwave Corporation
Inter-Logic Systems Company, Inc.
JFW Industries Inc.
Johanson Dielectrics, Inc.
Johanson Manufacturing Corporation
John Fluke Mfg. Co., Inc.
K&L Microwave, Inc.
Kalmus International
Kay Elemetrics Corporation
KDI Electronics
KW Microwave, Inc.
Kyocera America, Inc.
L & M Engineering
Litton Solid State
Loral Microwave — FSI
Lucas EPSCO Incorporated
M/A-COM/Adams Russell Electronics Co. Inc.
Magnum Microwave Corporation
Marconi Instruments
McDougall/McLaughlin — Mark V Associates
Menlo Industries
Merrimac Industries, Inc.
Metelics Corporation
Microwave Modules & Devices, Inc.
MSC (Microwave Semiconductor Corp.)
Microwave Solutions Inc.
Motorola Semiconductor Products Sector
Motorola Components Division
MSN/Microwave Systems News
Murata Erie North America, Inc.
Nedrud Data Systems
Pacific Monolithics
Phonon Corporation
Piezo Crystal Company
Q-Bit Corporation
Racal-Dana Instruments Inc.
Republic Electronics Corp.
RFD Inc.
RF Design
RF Monolithics, Inc.
RF Power Labs
Richardson Electronics, Ltd.
Sawtek Incorporated
Sciteq Electronics, Inc.
Semetex Microwave and RF Division
Signetics Corporation
Sprague-Goodman Electronics
Stanford Telecom
Stetco Inc.
Synergy Microwave
Tecdia, Inc.
Tech-Ceram Corporation
Techtrol Cyclonetics, Inc.
Tektronix Inc.
Teledyne Microwave
Tele-Tech Corporation
Telonic Berkeley, Inc.
Texas Instruments
Thomsonics
Toshiba America
Trilithic Inc.
TriQuint Semiconductor
TRM, Inc.
Trontech, Inc.
TTE, Incorporated
Unitrode
U.S. Instrument Rentals
Varian EIMAC
Vectron Laboratories, Inc.
Voltronics Corporation
Watkins-Johnson Company
Wavetek Corporation
Webb Laboratories
West-Bond Inc.
Wiltron Company

Cable Prep Tool

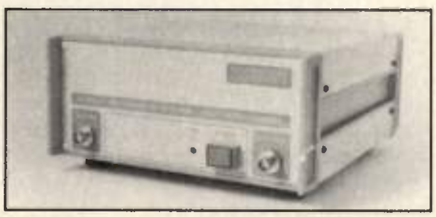
The EASIAX™ cable preparation tools are designed for use with HELIAX® cable. Each cut is made at the crest of the corrugation, at the required setback distance. The cut makes flaring the conductor easy, and the fixed blade depth eliminates accidental cutting of the inner conductor. Model 207865 is for 1/4 in. and 1/2 in. super-flexible HELIAX cable and Model 207866 is for 1/2 in. HELIAX coaxial cable. Both models feature reversible, off-center cutting blades to provide built-in spare cutting edges. **Andrew Corp., Orland Park, IL. INFO/CARD #226.**

Ruggedized Signal Generator

Wavetek RF Products introduces a 10 kHz to 1.1 GHz ruggedized signal generator. Model 2410R is a general-purpose unit that is suited for remote location communications testing applications. Features include an output level of −127 dBm to +13 dB, internal and external amplitude and frequency modulation with 400 Hz and 1100 MHz internal modulation sources, and an IEEE-488 interface. In single unit quantities, the signal generator is priced at $5995. **Wavetek RF Products, Inc., Indianapolis, IN. INFO/CARD #225.**

RF Power Amplifier

Model RC1001-1 is a multi-octave RF power amplifier with a 1 watt output from 100 kHz to 1000 MHz. Power gain is 30 dB and harmonics are −23 dBc. Input



and output impedance is 50 ohms and the amplifier operates from −10 degrees C to +45 degrees C. **Wessex Electronics Limited, Bristol, England. Please circle INFO/CARD #224.**

Switchable Filter Banks

Synergy introduces a new product line of signal processing modules which include switchable filter networks. In these subassemblies, one filter can be selected at a time from a number of filter types (lowpass, highpass, bandpass). The modules come in shielded connectorized packages and can accept frequencies from DC to 1000 MHz. Internal TTL drives are included.

Also from Synergy is a frequency converter module that translates incoming RF signals from DC to 2 GHz into customer-specified frequencies and levels. The shielded, connectorized package utilizes PC plug-in devices such as power dividers, double balanced mixers, frequency doublers, directional couplers, filters, attenuators and amplifiers. **Synergy Microwave Corp., Paterson, NJ. INFO/CARD #223.**

Spectrum Analyzer

The HP 3588A spectrum analyzer covers the 10 Hz to 150 MHz frequency range with a wide variety of frequency spans and resolution bandwidth settings from 20 kHz to 0.0045 Hz. It features a narrowband-zoom mode which uses an implementation of the FFT to provide

spectrum measurements in spans of 40 kHz or less. The narrowband-zoom span can be located anywhere within the 150 MHz span of the analyzer. Measurements in this mode are from 50 to 400 times faster than swept-tuned analyzers of comparable resolution bandwidths. All test results and measurement states can be stored and recalled either from internal non-volatile memory or a built-in 3 1/2 in. disk drive. The instrument is priced at $20,000. **Hewlett-Packard Company, Palo Alto, CA. Please circle INFO/CARD #222.**

Low Cost Frequency Synthesizer

Z-Communications introduces the S-700 Series of frequency synthesizers from 200 MHz to 2800 MHz with a switching speed of under 1 ms. Step sizes of 125 kHz, 250 kHz and 1 MHz can be obtained by externally programming the programmable reference divider to set up the correct division ratio for the on-board reference used. The on-board reference is provided with stability options to ± 5 PPM. Other features include phase noise of −90 dBc/Hz at 10 kHz offset from the carrier and a power output of +10 ±2 dB Harmonics are under −20 dBc and spurious is less than −60 dBc. In sample quantities, the S-700 is priced at $350. **Z-Communications, Inc., Ft. Lauderdale, FL. INFO/CARD #221.**

High Power Pulsed Transmitter

A line of high power pulsed transmitters which cover the 100 to 225 MHz, 225 to 500 MHz, 100 to 500 MHz, and 500 to 1000 MHz frequency range are being introduced by SPC. Specifications for the P/N SSPA-0510-100 P, 500 to 1000 MHz unit, include peak RF power of 100 watts, minimum instantaneous bandwidth of 700 MHz, maximum gain flatness of ±2 dB, input/output VSWR of 1.5:1 max., and spurious signals at −45 dBc. The unit exhibits gated Class AB operation. **System Planning Corp., Arlington, VA. INFO/CARD #220.**

Spectrum Analyzer/Tracking Generator

The HM 8028 is a spectrum analyzer that covers the 0.5 to 500 MHz (−3 dB) range. An oscilloscope in X/Y mode is used as a display with average noise level of −99 dBm. In addition to swept-tuned frequency mode, it can be used

INFO/CARD 55

in the fixed-tuned mode (zero scan) to provide time-domain measurement capability.

Also available is the HM 8038 tracking generator suitable for making wide dynamic range frequency response measurements on filters and amplifiers. When the two units are used together, they form a scalar network analyzer able to analyze frequency characteristics of various two-port devices such as active and passive filters, amplifiers and attenuators. The HM 8028 is priced at $768 and the HM 8038 is priced at $428. **Hameg, Inc., Port Washington, NY. INFO/CARD #219.**

Line Filters

Schaffner introduces the FP 720 family of line filters from 100 kHz to 1 GHz. The filters feature extended attenuation characteristics due to the use of ceramic feed-through capacitors for

the input and output connections. Internally, the filters use rod core chokes that attenuate transients and help limit the inrush of current. They are available with varistor/gas tube surge arrestor, providing protection against high voltage transients caused by lightning or nuclear electromagnetic pulses. The filters are available with current ratings from 1 to 16 amps. **Schaffner EMC, Inc., Union, NJ. INFO/CARD #218.**

RF Probe

The P-20 RF probe provides a 500 ohm input impedance to the circuit being tested, which is a sufficiently high value to minimize loading efforts in most RF circuits. Its construction produces less than 1 pF of input capacitance to

reduce circuit detuning. The probe features a 100 kHz to 2 GHz frequency response, 20 dB voltage attenuation (10:1) and an internal DC block (50 volts). The probe is priced at $49. **Auburn Technology Corp., Auburn, KS. INFO/CARD #217.**

Variable Center Frequency Filter

The Model CQT 1000 is a variable center frequency-stepped bandwidth filter with a 200 to 1000 MHz range. The bandwidth can be stepped from 1 percent to 5 percent (3 dB BW) in steps of 1 percent, 1.5 percent, 2 percent, 3 percent, 4 percent and 5 percent without loss of the nominal Butterworth shape factor. It is tunable over 2 1/4 octaves. VSWR is 1.5:1, and out of band rejection is 100 dB. **CQT Electronics, Inc., Beltsville, MD. INFO/CARD #215.**

Surface-Mount Chip Capacitors

ATC introduces a multilayer chip capacitor which offers the user the ability to adjust capacitance. STA-TUNE™ provides the user with the advantage of in-circuit adjustment without some of the disadvantages of trimmer capacitors. Twenty-two different values from 1 pF to 68 pF with adjustment ranges of 0.15 pF to 3 pF are available. Stable in-circuit adjustment is accomplished by connecting one of six different capacitance combinations. The capacitor measures 0.1·0 in. X 0.110 in. with a maximum height of 0.102 in. (case B). **American Technical Ceramics Corp., Huntington Station, NY. Please circle INFO/CARD #216.**

HF Whip Antenna

Astron introduces the VDW-153 high power (1 kW) HF whip antenna that is designed for operation from 1.5 to 30 MHz without the use of a coupler or tuner unit. The 35 foot whip is composed of six interlocking sections, and is constructed for quick disconnect and storage. The antenna has a maximum VSWR of 3:1. **Astron Corp., Herndon, VA. INFO/CARD #214.**

RF Pulse Amplifier

The 3130 Series amplifier covers from 200 to 500 MHz at power levels from 50 to 300 watts. Full power pulses are available for up to 250 ms. Standard

features include pulse droop of less than 5 percent, and blanking of under 2 us. The amplifiers are rack mount and operate on a variety of line voltages. **American Microwave Technology, Inc., Fullerton, CA. INFO/CARD #213.**

2 GHz Signal Generator

The Rohde & Schwarz Model SMGU covers a 100 kHz to 2.16 GHz frequency range without the use of doublers. Frequency resolution is 0.1 Hz, phase noise is −144 dBc/Hz at 100 MHz with 20 kHz offset, and residual FM at 500 MHz is less than 0.5 Hz. Output levels are available from −140 to 16 dBm. The instrument is priced at $25,800. **Rohde & Schwarz, Inc., Lanham, MD. Please circle INFO/CARD #212.**

Shielded Air Vent Filters

Spira unveils three classes of EMI/RFI shielded air vent honeycomb filters. These are commercial, military and military dual-panel assemblies. The commercial grade comes in three sizes of cell width and thickness and is designed to provide a relatively high level of shielding. The military grade has tin-plated honeycomb, frame and all joint surfaces to provide a high level of shielding and to survive the military environment over extended periods of time. The military grade comes in six varieties of cell widths and thicknesses. The military dual-panel filter adds a significant level of shielding, providing as much as 120 dB of shielding to E-fields at 1 GHz. **Spira Manufacturing Corp., North Hollywood, CA. Please circle INFO/CARD #211.**

Broadband Receiving Antenna

The SAS-2/A is a broadband receiving antenna for measuring electric field intensity at frequencies from 100 Hz to 960 MHz in a single band. Low-noise amplifiers allow the detection of E-fields as weak as −130 dB (V/m) over most of the band. A general roll-off in the response below 10 kHz prevents overload due to strong 60 Hz harmonics. The antenna is intended primarily for use indoors, including shielded rooms, but can also be used outdoors if properly protected from the weather. **Antenna Research Associates Inc., Beltsville, MD. INFO/CARD #210.**

Spectrum Probe

Smith Design introduces the 107 Spectrum Probe™ as a scope accessory which produces a spectrum analyzer

display. A 10 pF, 500 VDC capacitor isolates the input stage, allowing transparent exploration of circuit operation. Specifications include a frequency range of 1 to 100 MHz and a dynamic range greater than 50 dB. The scope bandwidth requirement is 500 kHz. List price is $380. **Smith Design, Dresher, PA. INFO/CARD #209.**

Surface-Mount Amplifiers

This thin-film, cascadable 20 to 400 MHz amplifier series has +16 dBm of output power, 14.4 dB gain and 3.7 dB noise figure. Third order and second order intercept points are +32 dBm and +44 dBm, respectively. The PPA-441 is packaged in the Avantek PlanarPak™ surface-mount package which measures 0.375 inches square and weighs 0.5 grams. The package is designed for use on a single-sided, microstrip-based motherboard. **Avantek, Inc., Milpitas, CA. INFO/CARD #208.**

New Cable Assembly

Northeast Coax introduces a cable assembly that uses two crimp style SMA connectors and RG 316 cable. The frequency range for this assembly is DC to 10 GHz. VSWR is approximately 1.18 + 0.024f(GHz) and insertion loss in dB is 0.045(f)1/2(length in inches). **Northeast Coax, Hamden, CT. Please circle INFO/CARD #207.**

rf software

Filter Design Software

This program is capable of designing LC bandpass filters from 10 Hz to 2 GHz with up to 20 coils. The lossy bandwidth is automatically adjusted to an accuracy better than 0.1 percent using components that have practical Q values (specified). The program designs Butterworth, Chebyshev, Bessel and Constant-K bandpass filters having coils equal to a specified optimum inductance value, and narrowband and Norton transformations are performed automatically. The program, is priced at $99 postpaid. **Phil Geffe's Filterwave, Cincinnati, OH. INFO/CARD #197.**

S-Parameter Transfer Program

Netcom automates the process of transferring S-parameter data from a network analyzer to CAD programs such as RoMPHE, Microwave Harmonica, SuperCompact and CADec 4. The data collected can be used to generate accurate linear and non-linear models for RF, microwave and millimeter-wave circuit simulation, or can be used to develop a statistical database for Monte Carlo analysis and yield optimization. Any stored file can be read and displayed in both rectangular and Smith chart forms. Up to three traces can be overlayed for comparing different devices or different bias conditions. It is available for IBM PC, AT and compatibles with EGA. **Compact Software, Inc., Paterson, NJ. INFO/CARD #196.**

EMI/RFI Equipment Software

Electro-Metrics has developed software for use with their EMI and RFI test equipment to provide users with increased flexibility and accuracy. It controls test receivers via an AT-compatible computer equipped with an IEEE-488 interface card. The software operates using three main programs. The Create-A-Test routine loads all the information necessary to run a test including all receiver settings. The Scan routine controls the receiver and collects data, and the Zoom routine transfers the program control to the zoom routine, offering the user static zoom or active zoom analysis tools. **Electro-Metrics, Amsterdam, NY. INFO/CARD #195.**

rf literature

Filter Catalog

TTE introduces a catalog which features a line of active and passive filters which includes Bessel, Butterworth, Chebyshev, elliptical function, Gaussian, programmable, notch, and custom designs. The filters are available for frequencies from 0.1 Hz to 500 MHz. Normalized test data and response curves covering the passband, return loss, time delay, phase linearity and overall response information are included. **TTE, Inc., Los Angeles, CA. INFO/CARD #194.**

Coaxial Cable Phase Characteristics Data Sheet

This technical summary provides detailed information on the phase characteristics of Andrew Corporation's Heliax® coaxial cable. Phase change data as a function of both temperature and bending are provided, including graphs showing phase response from −50 degrees C to +80 degrees C. A comparison of flexible cable to semi-rigid and braided cable is included. **Andrew Corporation, Orland Park, IL. INFO/CARD #193.**

Bulletin Features Ceramic Dielectric Trimmer Capacitors

Sprague-Goodman is offering an engineering bulletin that features their line of ceramic dielectric trimmer capacitors designed for standard and surface-mount applications. It lists features, specifications, standard rating charts, schematic drawings and application notes. Gull-Wing versions of the Surftrim® line of surface-mounted trimmers, together with carrier and reel specifications, are highlighted. Also shown are ruggedized 5 mm and 7 mm types, plastic encased 5 mm and 4 X 4.5 mm types, plus miniature and microwave trimmers for hybrid circuit applications. **Sprague-Goodman Electronics, Inc., Garden City Park, NY. Please circle INFO/CARD #192.**

Book Discusses Network Theorems

Useful Network Theorems With Applications by Dr. Harry E. Stockman discusses modern network theorems. These include Helmholtz', Mayers's, and Katzimierczuk's theorems, as well as several new theorems contributed by the author. Over 100 problems are solved in this book. It is priced at $11.25. **Sercolab, E. Dennis, MA. Please circle INFO/CARD #191.**

High Performance Linear Circuits Guide

Burr-Brown announces the availability of a guide which features technical abstracts on over 70 high-performance, high-quality linear products. These include data converters, operational and instrumentation amplifiers, power amplifiers, isolation amplifiers, sample/hold amplifiers, voltage-to-frequency converters, DC/DC converters, signal conditioning modules, and digital signal processing products. Information on requesting technical assistance is also included. **Burr-Brown Corporation, Tucson, AZ. INFO/CARD #190.**

Quartz Crystal Brochure

This brochure from Electro Dynamics Crystal Corporation includes specifications on their line of military quartz crystals. A tutorial section is featured which highlights operational theory and other topics of interest such as the crystal equivalent circuit, impedance graph, temperature coefficients, and related engineering formulae. Also described are application and ordering information and design engineering services. **Electro Dynamics Crystal Corporation, Overland Park, KS. Please circle INFO/CARD #189.**

Noise Figure Application Note

Eaton Corp. announces the availability of an application note entitled "Measurements of the Noise Figure of Amplifiers, Mixer-Amplifiers and the Use of Filters in the Measurements Set-Up." The note deals with uncertainties present when measuring mixer-amplifiers with image frequencies or systems with narrow band filters. This theoretical analysis is complete with descriptive figures and equations. **Eaton Corp., Electronic Instrumentation Division, Los Angeles, CA. INFO/CARD #188.**

Diode Catalog

Teklec Microwave introduces a silicon microwave diode catalog which features PIN diodes including high voltage units up to 3 kV, Schottky barrier diodes, tuning varactor diodes, and power generation diodes. The devices are supplied for commercial, military and space applications. **Thomson Electron Tubes and Devices Corp., Totowa, NJ. Please circle INFO/CARD #187.**

Spice Newsletter

The intusoft Spice newsletter discusses analog circuit simulation and related information on various types of electronic devices, circuits, Spice models, and summation techniques. It also provides application notes. The current issue of this bimonthly publication discusses a Spice model for a transimpedance operational amplifier. **intusoft, San Pedro, CA. INFO/CARD #186.**

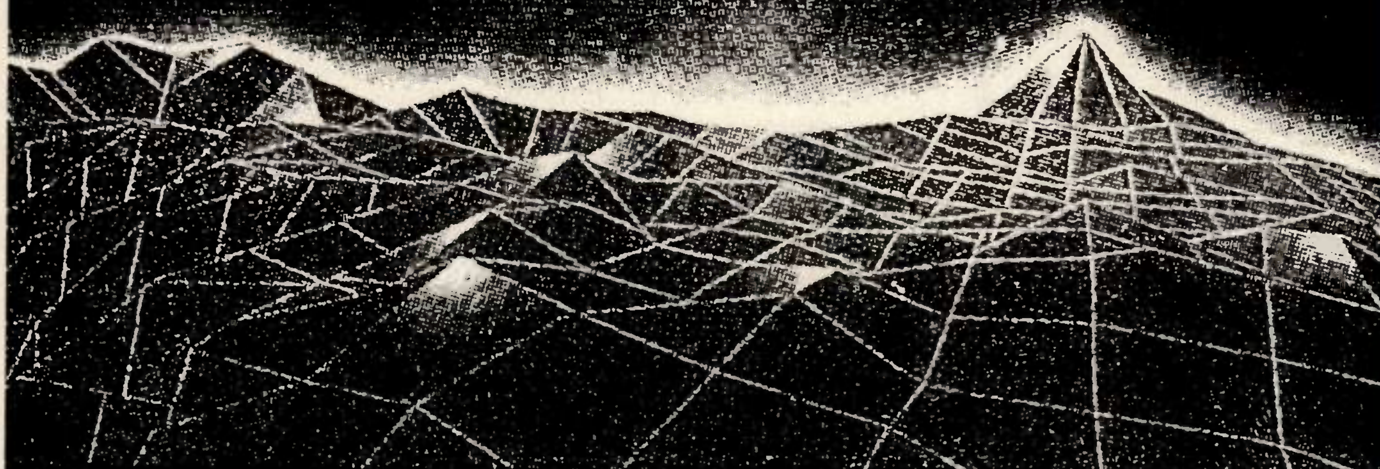
MMIC Screening Brochure

This brochure from CEL details screening and qualification procedures for NEC silicon MMICs, GaAs MMICs and high-speed digital GaAs ICs. Circuit elements of the NEC devices covered in the brochure are qualified under MIL-STD-883. Reliability data is provided for both chips and packages for JAN Class S equivalent, JAN Class B equivalent, JANTX equivalent and JAN equivalent. The processing information and flow diagram provided make it easy to select reliability grades and complete specification control drawings. **California Eastern Laboratories, Santa Clara, CA. INFO/CARD #185.**

Military Products Brochure

SAW devices, SAW-stabilized frequency sources and integrated SAW products are covered in this brochure from RF Monolithics (RFM). Products discussed include SAW resonators and coupled-resonator filters up to 1650 MHz, low-loss and conventional SAW transversal filters up to 850 MHz, and low-loss and wideband delay lines up to 1600 MHz. A review of RFM's manufacturing and quality capabilities for military products, plus recommended screening and product qualification programs for MIL-S-49433 and other military standards, is included. **RF Monolithics, Inc., Dallas, TX. INFO/CARD #184.**

CHANGE YOUR PERSPECTIVE



Setting milestones in the cellular market is par for the course at Motorola. Our environment stretches the imagination, encourages "zero defects" and gives **ENGINEERS** a new perspective on the realm of technological possibilities.

Winning the Malcolm Baldrige Quality Award in 1988 is tangible proof how we reach our goals. If you would like to challenge your perspective and your future in cellular technology, take a look at the following opportunities:

SOFTWARE: BSEE/BSCS/BSCE or equivalent. Experience with real-time software development using "C", UNIX*, Pascal and structured design; CASE software tools (Sun, Apollo); design, programming and system debugging for telephone digital switching or data communications systems; assembler and high-level structured languages; hardware diagnostics; object-oriented design, C++; software development tools such as Teamwork and Interleaf; ASN.1 (Abstract Syntax Notation); data communications protocols (CCITT #7, X.25, X.75, PAD, LAPB, LAPD, Ethernet, ISDN, CCITT V Series modem, group 3 fax); modeling and simulations (GPSS, SIMSCRIPT and other simulation tools) desired. **Refer to Dept. #SE/YF.**

ELECTRICAL/COMPUTER: BSEE/MSEE/BSCE/BSCS or equivalent. Seeking experience in any of the following areas: analog and RF circuit design, receiver or transmitter RF circuits; LCD display technology, acoustics, and 8-bit microprocessor software; infrared and fiber optics; low speed data; digital modulation; 900 Mhz RF power amplifier design with hybrid or microstrip circuitry; RF device development; parallel and/or push-pull RF amplifier design at 900 Mhz or UHF; A/D and D/A convertors; RF propagation; passive/active filter theory; microwave antenna design; UL, EMI and RFI requirements; HP 3062 test system; HP 9000; digital hardware, microprocessor applications and interfaces; digital switching technology; firmware development methodology; PCM and digital telephony; digital signal processing; ASIC design; LAN systems, PSTN standards, ISDN standards, processor architectures, high speed logic; 16-bit MPU design practices, programmable logic; digital modulation, synchronization, adaptive signal processing, viterbi algorithm and MLSE, decision feedback equalizer; convolutional code and linear block code, speed coding, echo cancellation, encryption; CAE techniques for digital hardware design using Mentor Graphics, etc.; data communications protocols (CCITT #7, OSI, X.25, X.75, PAD, LAPB, LAPD, Ethernet, ISDN, CCITT V Series modem, group 3 fax); computer network management/administration (Apollo, Sun, Mentor Graphics, AppleTalk). **Refer to Dept. #EC/YF.**

SYSTEMS: BSEE/BSCE/BSCS or equivalent. We are seeking persons with experience in any of the following technologies: radio/cellular communication systems engineering; RF propagation prediction methods; PSTN traffic planning; telephone network, interconnection and telecommunication industry standards; microwave system design and equipment engineering; telephone switching systems; software programming skills. **Refer to Dept. #SY/YF.**

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