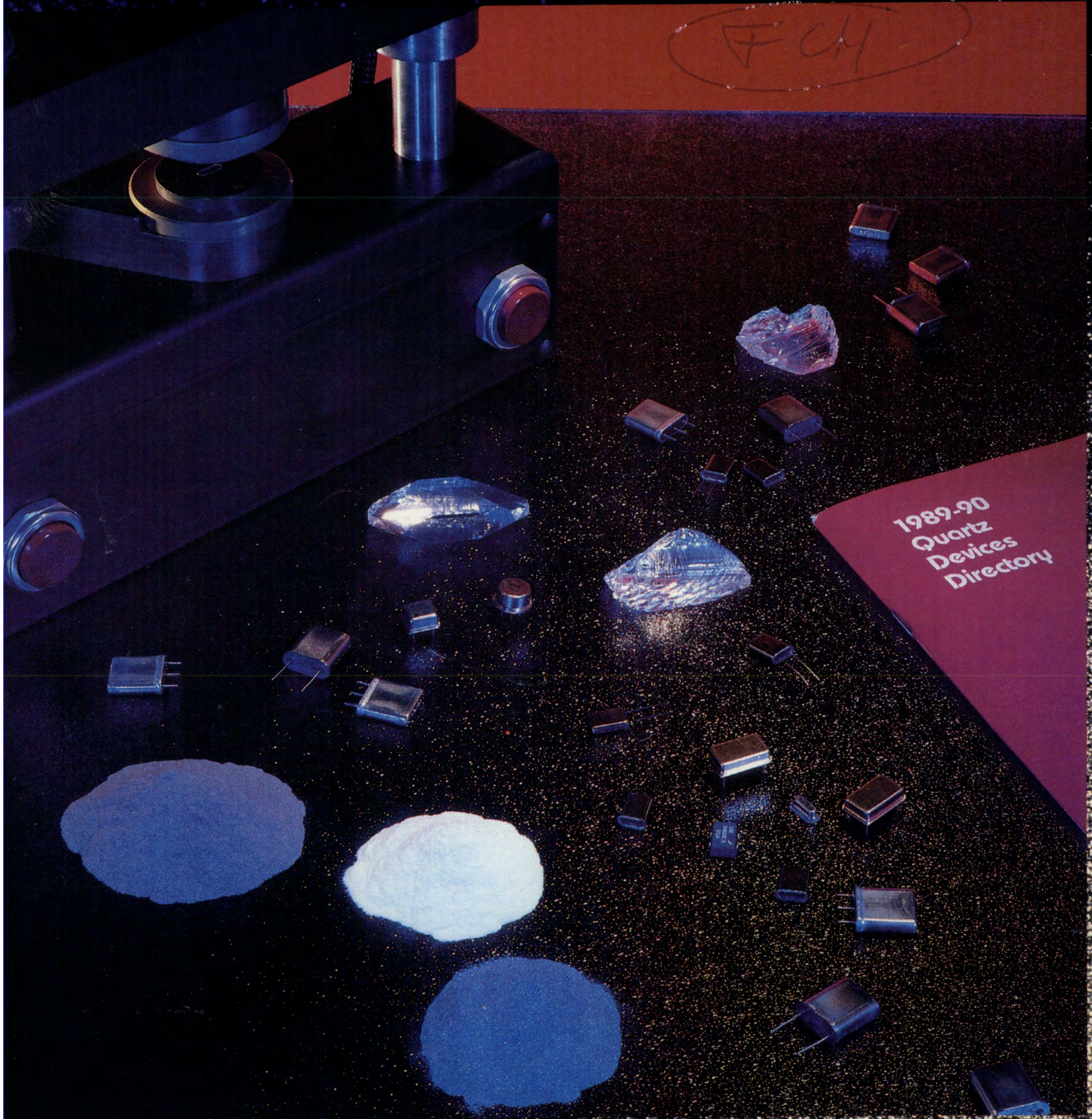
RF design

engineering principles and practices

September 1990

FCH



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12th Piezoelectric Devices Conference

industry insight

RF Integrated Circuits

$$t_{Rise/Fall} = 0.0000000007$$

$$t_{Rise/Fall} = 0.7\text{x}10^{-9} \text{ seconds}$$

$$t_{Rise/Fall} = 0.7 \text{ nanoseconds}$$

$$t_{Rise/Fall} = 700 \text{ picoseconds}$$

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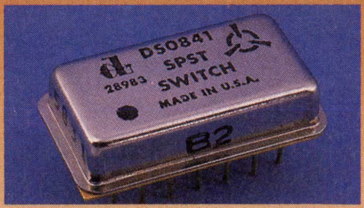

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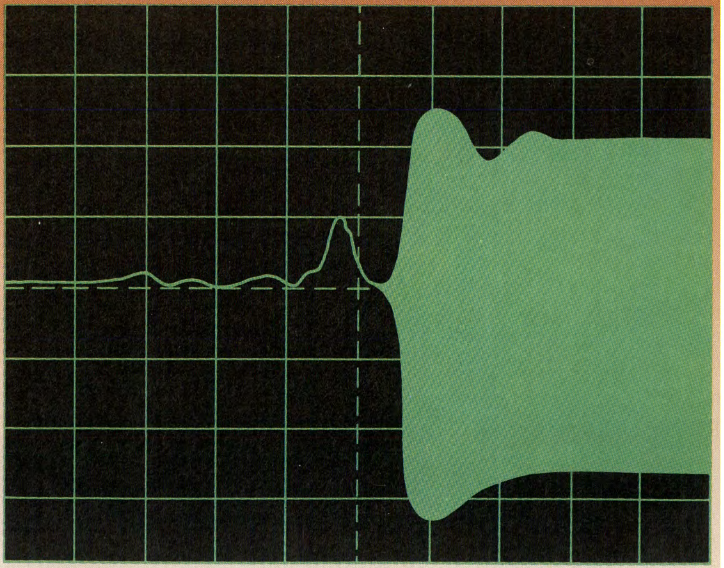
DSO841 also offers excellent isolation of 70dB typical at 100MHz typical and 60dB typical to 200MHz.

The DSO841 is the ideal high speed, low transient, low power consumption switch for blanking and modulating applications.

Operating Characteristics

PARAMETER	MIN	TYP	MAX	UNITS	CONDITIONS
Current Drain		2	5	mA	AT +5V DC Supply
Switching Transients		56	100	mV	Peak Value
Transition Time		.7	1	nS	90%/10% or 10%/90% RF
Switching Speed		7	10	nS	50% TTL to 90/10% RF
Insertion Loss		1.7	2.3	dB	
Isolation	60	70		dB	10-100
	50	60		dB	100-200
Operating Frequency	10		200	MHz	
TTL Controlled					



Scope Specifications

Ch.1 = 100.00 mvolts/div Timebase = 1.00 nsec/div Offset = 20.00 mvolts Delta T = 480.0

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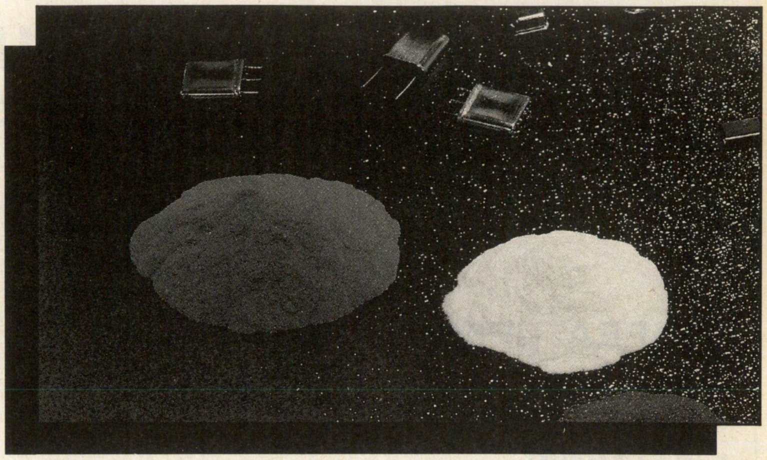

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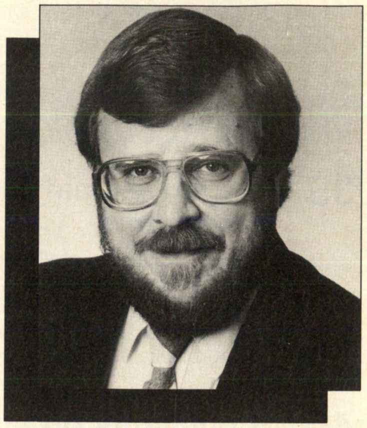
RF editorial

Computers and RF Engineering



By Gary A. Breed
Editor

Our industry has been fairly slow in reaching the point of "a computer on every desk," something that seems unusual for a high-tech profession. Just a few years ago in 1986, our reader survey found that 11 percent of engineers had to wait in line to use shared services. Although 74 percent had access to a PC at that time, nearly 60 percent still relied on a programmable calculator for much of their work.

The most telling point in comparing then with now is the type of PC used. In 1986, over 27 percent of engineers used something other than an MS-DOS compatible machine. Of that group, Radio Shack models were mentioned twice as many times as Apple!

Since then, we have been paying close attention to the use of computers in the RF engineering design process — and things have certainly changed. Now, most of the letters we receive are dot-matrix printed, not typewritten. We get enough articles with accompanying computer programs that we started the RF Design Software Service to distribute them (it has been a well-used service, too). It's our conclusion that RF engineers have pretty much replaced their HP-41 and TI-59 calculators with PCs.

In this issue, our monthly survey is on RF software. Fill it out and let us know how you currently use computers and RF software products. We can then update some of the information gathered in 1986.

Now that the power of computers is being used by most RF engineers, we think that the best of the group should earn some well-deserved recognition. Last month, we announced the new PC Software category in our annual RF Design Awards Contest. Some terrific programs will win some great prizes. Two years ago, we wouldn't have thought to have a contest for RF software, but it seems perfectly obvious now.

Of course, the original circuit design contest is still here, having continually gotten stronger over its five-year history. We are just doubling the fun with another entry category. The top prizes for both categories will be announced in the next issue, and complete rules were published on page 50 of the August issue. Take a look, and get to work on your entry. You have until next March 30, but don't procrastinate.

As a final note on the subject, I've noticed that most RF engineers regard computers differently than the "digital" types, treating them as tools, not toys. They get *used* to assist in the design process, just like spectrum analyzers, reference books, reflection coefficient bridges, and soldering irons. I used to be worried that computers would create engineers who would use them to create circuits without really understanding the design process. Now that computers are universal, and I have seen much of the software available for design assistance, those worries are gone. Clever use of software might cover up an engineer's incompetence for a while, but I know that he can't hide there for long.

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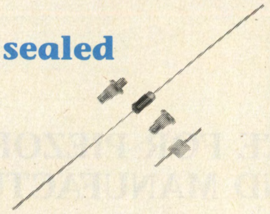

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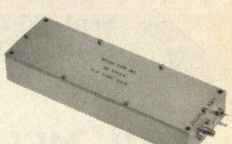

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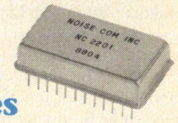

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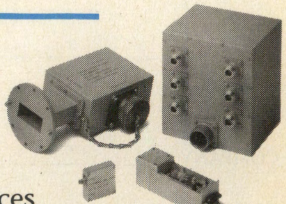

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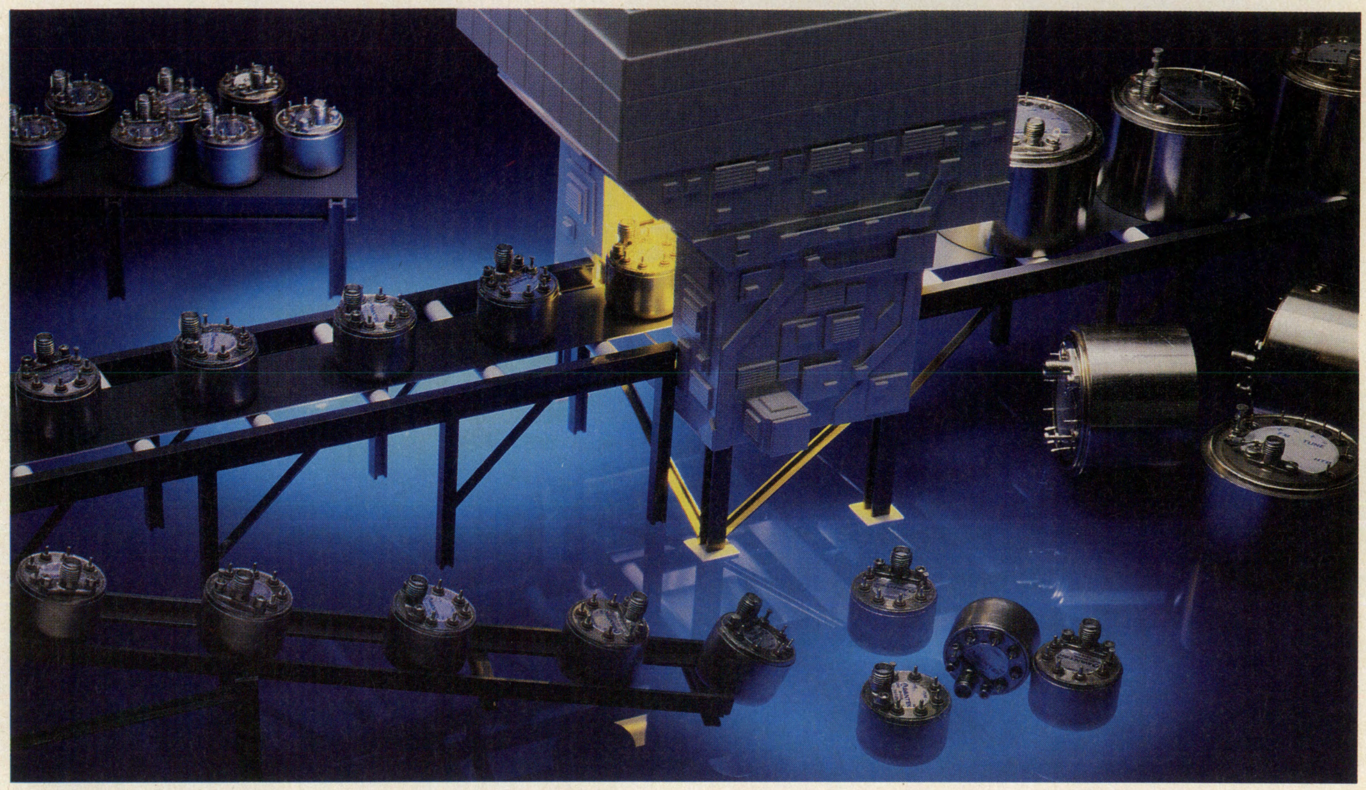

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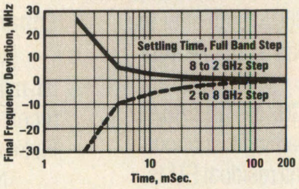

AV-7036 covers the 3.0 to 6.0 GHz range. AV-7028 provides 2.0 to 8.0 GHz frequency coverage. For both models, phase noise is typically −100 dBc/Hz at 20 kHz offset from carrier. And, power output is +13 dBm.

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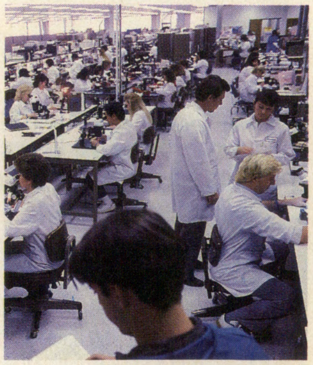

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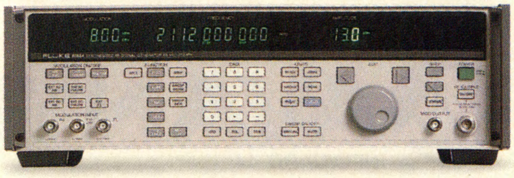

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SSB Phase Noise @1 GHz			
20 kHz offset	-131 dBc/Hz	-134 dBc/Hz	-128 dBc/Hz
1 kHz offset	-94 dBc/Hz	No Specification	-91 dBc/Hz
Non-harmonic Spurious @1 GHz	-100 dBc	-100 dBc	-100 dBc
Residual FM @1 GHz	1.5 Hz	2 Hz	2 Hz
Modulation	AM, FM, øM, Pulse	AM, FM, øM, Pulse	AM, FM, øM, Pulse
Pulsed RF			
Rise/fall time	15 ns	400 ns	100 ns
On/off ratio	80 dB	40 dB to 80 dB	35 dB to 80 dB
Programming Compatibility	HP 8642A/B, Fluke 6060, 6070	HP 8642A/B	HPSL
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Letters should be addressed to Editor, *RF Design,* 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111.

Transmission Line Matching Circuits

Editor:

In reference to the following:

"Design of Transmission Line Matching Circuits" by Stanislaw Rosloniec, *RF Design,* February 1990, Reference 5 and Letter to the Editor from James J. Lev, *RF Design,* June 1990. The above subjects all deal with impedance matching by a single lossless transmission line; and it seems that now three people claim authorship of the basic formulas, given as equations 1 and 2 in Rosloniec's paper. So the time has perhaps come to point out that equation 2, which is identical to equation 8 in Reference 5, contains a sign error. The correct formula for the electrical length of the matching transmission line is

$$\theta = \arctan (R_1 - R_2 / R_1X_2 + R_2X_1)Z_0$$

In other words, both terms in the denominator are positive. This can easily be shown by solving the basic equation for impedance transformation by a transmission line of characteristic impedance Z_0 and electrical length θ having a load impedance $Z_2 = R_2 + jX_2$ and input impedance $R_1 + jX_1$:

$$R_1 + jX_1 = [Z_2 + jZ_0 \tan\theta]/[1 + j\,(Z_2/Z_0) \tan\theta]$$

Rosloniec's example, matching circuit I of Table 1, assumes a load impedance of $R_2 + jX_2 = (30 - j10)\Omega$ and a desired input impedance $R1{+}jX1 = (50 + j20)\Omega$. The transforming line needs a characteristic impedance $Z_0 = 43.0\ \Omega$, but its length is not 2.478 rad, as stated, but only 1.46 rad. Referred to $Z_0 = 43\ \Omega$, the load and input reflection factors are

$$r_2 = 0.223 \measuredangle -135 \text{ degrees}$$
$$r_1 = 0.223 \measuredangle\ 58.6 \text{ degrees}$$

The transformation runs through 166.4 degrees on the Smith chart, which corresponds to 83.2 degrees electrical length.

As the final verification, the circuit was analyzed with Touchstone, using the following circuit file:

```
CKT
  TLIN 1 2 Z=43   E=83.2  f=1
  SRC 2 0 R=30    C=15.9
  Def1P 1 Line
OUT
  Line S11
FREQ
  Step 1
```

E is the electrical line length in degree. The load is modeled as a series connection of a 30 Ω resistance and a capacitance of 15.9 pF at 1 GHz (which amounts to a reactance of −10 Ω). At a frequency of 1GHz, the resulting magnitude of S_{11} was 0.196 and the angle of S_{11} was 78.730. This is the input reflection factor referenced to 50 Ω, which represents an impedance of exactly (50+j20)Ω. Hence, there is no doubt that the signs in the denominator of the equation for θ need to be positive.

Wolfgang Wiebach
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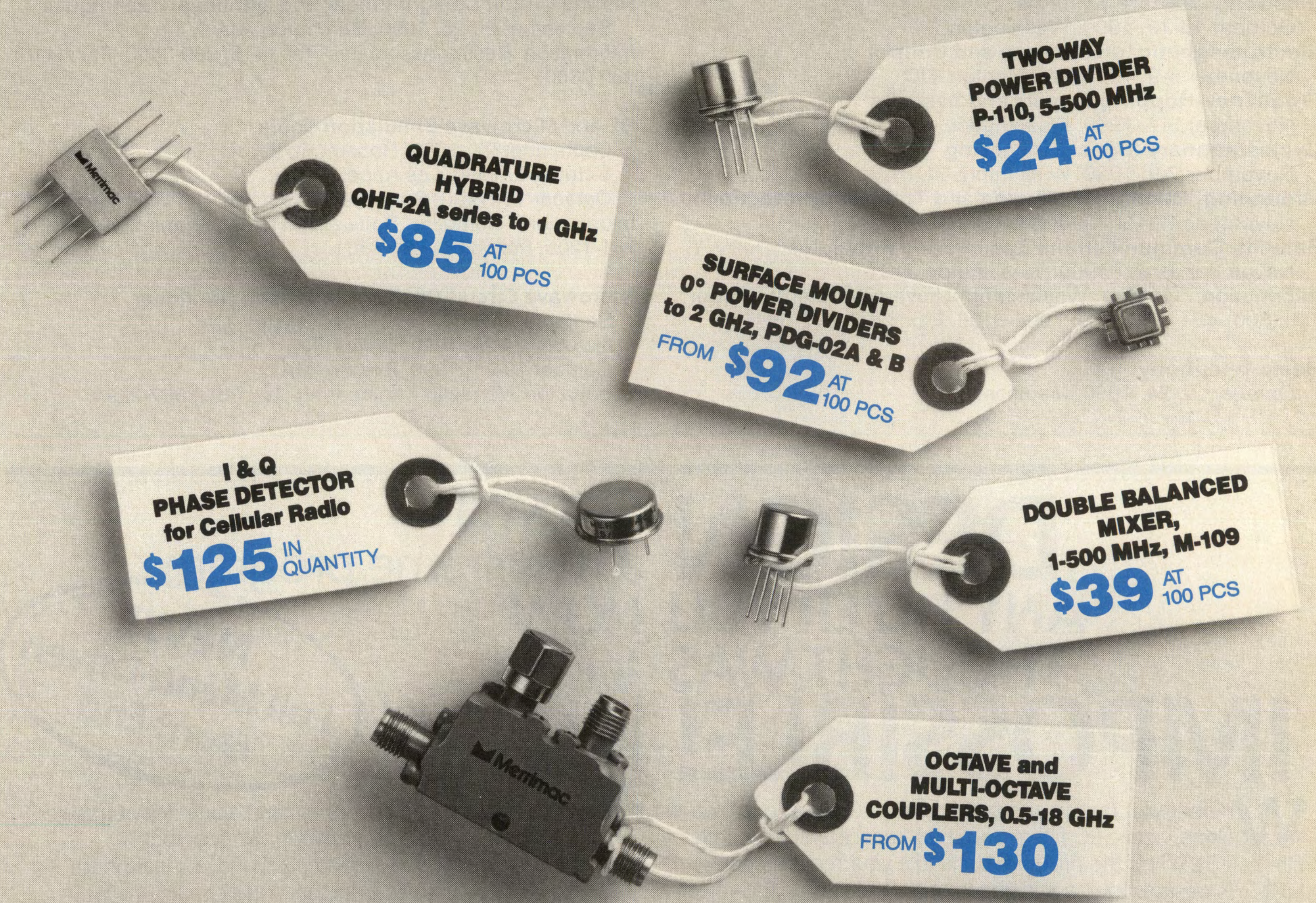

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Information: University of Wisconsin-Madison, College of Engineering. Tel: (800) 222-3623; (414) 227-3200. Fax: (414) 227-3119.

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September 24-28, 1990, Los Angeles, CA

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Information: Besser Associates. Tel: (415) 969-3400. Fax: (415) 965-0800.

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Information: Compact Software, Helen Shapiro. Tel: (201) 881-1200. Fax: (201) 881-8361.

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Information: Vendelin Engineering. Tel: (408) 867-2291.

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Design for ESD and RFI
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Information: The Keenan Corporation, Ms. Jean Whitney. Tel: (813) 544-2594. Fax: (813) 544-2597.

EMC for Digital Designers
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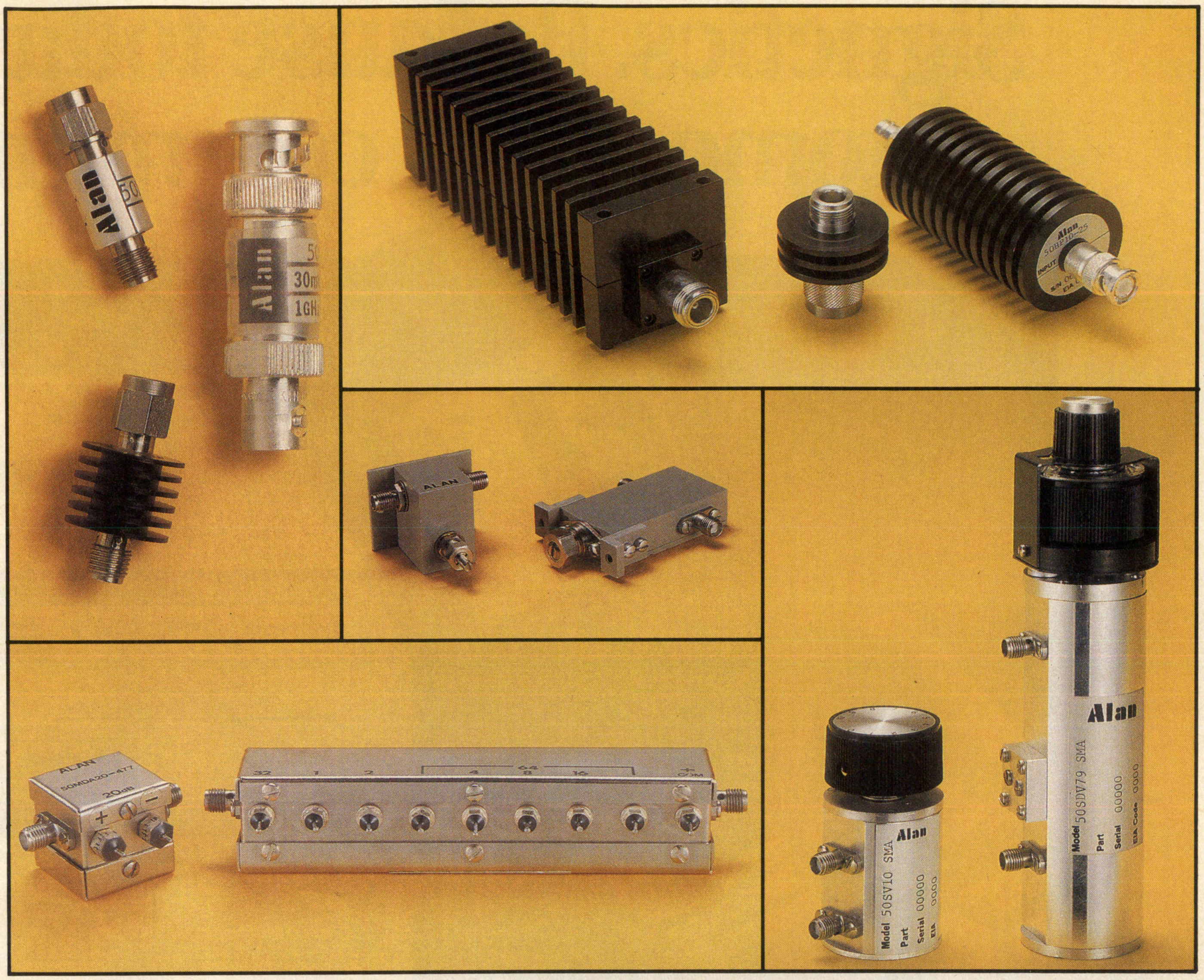

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25-27 **Piezoelectric Devices Conference**
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8-11 **Antenna Measurement Techniques Association Symposium**
Wyndham Franklin Plaza, Philadelphia, PA
Information: Ms. Jennifer Wentz, 1990 A.M.T.A. Symposium, Flam & Russell, Inc., P.O. Box 999, Horsham, PA 19044. Tel: (215) 674-5100.

15-18 **Electrical Manufacturing and Coil Winding '90**
O'Hare Expo Center, Chicago, IL
Information: Electrical Manufacturing & Coil Winding '90, 2400 East Devon Avenue, Suite 205, Des Plaines, Illinois, 60018.

17-20 **EMC Expo 90**
San Mateo County Expo Center, San Mateo, CA
Information: EMC Technology, P.O. Box D, State Route 625, Gainesville, VA 22065. Tel: (703) 347-0030.

30-Nov 1 **Electronic Imaging Conference East**
Hynes Convention Center, Boston, MA
Information: MG Expositions Group, 1050 Commonwealth Ave., Boston, MA 02215. Tel: (800) 223-7126 or (617) 232-3976.

November

6-8 **22nd International SAMPE Technical Conference**
Boston Park Plaza Hotel, Boston, MA
Information: SAMPE, P.O. Box 2459, Covina, CA 91722. Tel: (818) 331-0616.

13-15 **RF Expo East 90**
Marriott Orlando World Center, Orlando, FL
Information: Kristin Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600; (800) 525-9154.

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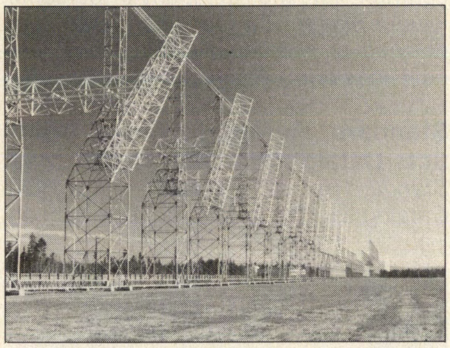
Over The Horizon Radar Turned Over to Air Force

The East Coast Over-The-Horizon (OTH) Backscatter radar site in Maine was recently turned over to the Air Force Electronic Systems Division by the main contractor, General Electric's Aerospace System Division. The Air Force has begun an operational test and evaluation of the system and will transfer it to the final user, Tactical Air Command, next year. The system in Maine covers a 180 degree span from northeast Canada and Greenland to Cuba at ranges from 500 to 1,800 nautical miles. The system has the ability to track aircraft of all sizes from bombers down to small, private planes. OTH bounces



its transmitted and received signals off the ionosphere thereby allowing the radar to detect aircraft at all altitudes from the ionosphere down to the earth's surface. While normal radars are limited in their range because of the curvature of the earth, OTH radar can cover up to 10 times the normal range which allows for increased coverage and warning time.

The East Cost OTH-B Radar System operations center is located at the Maine Air National Guard Base in Bangor, the transmitter site is near Moscow, Maine; and the receiver site is at Columbia Falls, Maine. Transmit and receive sites are separated to allow for continuous operation of the transmitters without interference in the receivers. At the transmit site, the coverage is divided into three 60 degree segments. Each segment is served by twelve 100 kW transmitters which deliver their signals into three 3,600 foot long antenna arrays. Return signals are detected at the Columbia Falls receiver site by one of three receiver antenna arrays.

The Maine site is the first of four OTH-B radar systems to be built. The West Coast system is nearing completion and construction on the Alaska system is slated to begin soon. However, there is some doubt as to the fate of the proposed Central System which would cover the Caribbean basin. In an effort to cut defense costs, the Senate recently recommended in their $289 billion Fiscal 1991 Defense Authorization bill, that the system not be built. The Senate's proposed bill would not affect the other three sites, as money for them was allocated from 1985 to 1987.

Improved Testing for EM Susceptibility — Two new NIST publications discuss improved methods for testing aircraft, large operation systems, and electronic equipment for susceptibility to electromagnetic fields. *Recent Improvements in Time-Domain EMC Measurement System* (NISTIR 89-3927) describes techniques for determining critical resonant frequencies and the current response of internal wiring of helicopters due to external EM fields. The measurement method uses a train of low-level radiated pulses that do not disturb other spectrum users, are safe, and can be used in a noisy EM environment. *Facilities for Improving Evaluations of Electromagnetic Susceptibilities of Weapon Systems and Electronic Equipment* (NISTIR 89-3928) discusses the preliminary design of a facility for EM susceptibility testing that combines features of the transverse electromagnetic cell for low frequency testing and the reverberating chamber for high frequency operation. Both publications are available from the National Technical Information Service, Springfield, VA 22161. Order NISTIR 89-3927 by PB #90-155821 for $15 and NISTIR 89-3928 by PB #90-155862 for $15.

Cubic Awarded Defense and Transit Contracts — Cubic Defense Systems' Aerospace Group received a $34.5 million contract for an instrumentation system from the U.S. Army Training and Doctrine Command. The Combat Maneuvering Training Center will track armored and ground forces in mid- to high-intensity combat. The aerospace group also received a $10.3 million contract for the company's personal locator system that locates crews from downed or missing aircraft behind enemy lines or other difficult situations.

Standard Reference Data Products Catalog Available — The *NIST Standard Reference Data Products 1990 Catalog* is now available. The catalog

INFO/CARD 20

INFO/CARD 21

provides the latest information on various data compilations, publications, and computerized databases that may be obtained from NIST and other sources. Data compilations for science and industry are available in the following areas: analytical chemistry, atomic physics, chemical kinetics, materials properties, molecular structure and spectroscopy, thermochemistry and the thermophysical properties of fluids. To obtain a copy of SP 782, send a self-addressed mailing label to Standard Reference Data Program, A323 Physics Bldg., NIST, Gaithersburg, MD 20899.

QUALCOMM and BOATRACS Enter Agreement — QUALCOMM recently announced that BOATRACS will be the exclusive distributors for OmniTRACS[R] - their two-way satellite communications and position reporting system. The agreement includes the sale of all OmniTRACS equipment and messaging/position location services for boats, ships and non-governmental maritime vessels located in domestic coastal waters of the Atlantic and Pacific Oceans. In launching this enterprise, BOATRACS will provide a total communications network capability connecting an offshore boat or ship with all appropriate shore-side services. These services include marine hardware outlets, fish markets, new crew notice boards, maintenance facilities and so forth.

Texas Instruments and DSO Announce Agreement — Texas Instruments and the DSO Group have announced a 10 year strategic agreement in which the DSO Group will develop and market chip sets for selected markets and customers based on TI's digital signal processing integrated circuit technology. The DSO Group will establish its own design centers to develop these semicustom chip sets. TI will manufacture the chip sets for them as well as provide technological support. Digital signal processing technology allows the manipulation of analog signals for speech, image, audio and radio transmission by digital computers for information extraction.

A.M.T.A. Conference Set for October — The Twelfth Annual Antenna Measurement Techniques Association Meeting and Symposium is set for October 8-11 in Philadelphia, PA at the Wyndham Franklin Plaza Hotel. This year's symposium is offering some excellent papers including: "Dielectric Materials Measurements," "Modern Diagnostic Evaluation of Absorber Materials,""A Cost Effective 50 dB Shield for Free Space Anechoic Chambers and Compact Ranges," "RCS Measurements in VHF/UHF in Indoor Microwave Range," "Calibrations of Large Antennas and Radio Stars," and "A New Concept for UHF/L-Band Compact Range Antenna Feeds." In addition, there are approximately 100 other papers on various topics slated for presentation. There will be approximately 16 exhibitors in attendance. For those interested in attending or obtaining more information, contact: Ms. Jennifer Wentz, 1990 AMTA Symposium, Flam & Russell, Inc., PO Box 999, Horsham, PA 19044. Tel: (215) 674-5100. Fax: (215) 674-5108.

Techalog Data Systems Ltd. Changes Name — Telecommunications Techniques Corporation has announced that its subsidiary in the United Kingdom, Techalog Data Systems Ltd., has changed its name to Telecommunications Techniques Company Ltd. The UK subsidiary sells and services its parent company's products.

AMP Acquires Kaptron — AMP Incorporated recently announced the acquisition of Kaptron, Inc., a manufacturer of passive fiber optic components, PC-based fiber optic test instrumentation, and manufacturing equipment. Details of the acquisition were not disclosed.

MFC to Acquire NSI — Microwave Filter Company and Niagara Scientific recently announced that they have agreed to merge. In the deal, MFC will acquire 100 percent of NSI capital stock in exchange for shares of MFC stock. NSI will become a wholly owned subsidiary and will operate as a separate company with the current management in place. NSI manufactures environmental monitoring equipment and industrial automation equipments. MFC manufactures electronic filters used for preventing interference or signal processing in cable television, satellite, broadcast, aerospace and government markets.

Instrumentation and Measurement Technology Conference Call for Papers — The IEEE Instrumentation and Measurement Technology Conference, to be held May 14-16, 1991 has issued a call for papers dealing with test instrumentation and measurement technology. Specific topics will fall in two categories: Measurement Techniques and Hardware - antenna/RCS/EM fields, sensors, integrated systems, transducers, standards and calibration, including self calibration, design of instrumentation, and parameter characterization of materials and devices; and Measurements, Automation and Computer Applications - expert systems and data reduction, ADCs and automation, software in instrumentation and measurement, and mixed analog and digital test techniques. The deadline for a 200 to 300 word abstract is October 8, 1990. Abstracts should be submitted to Conference Coordinator Robert Myers, c/o Myers/Smith Inc., 3685 Motor Ave., Suite 240, Los Angeles, CA 90034. Fax: (213) 287-1851.

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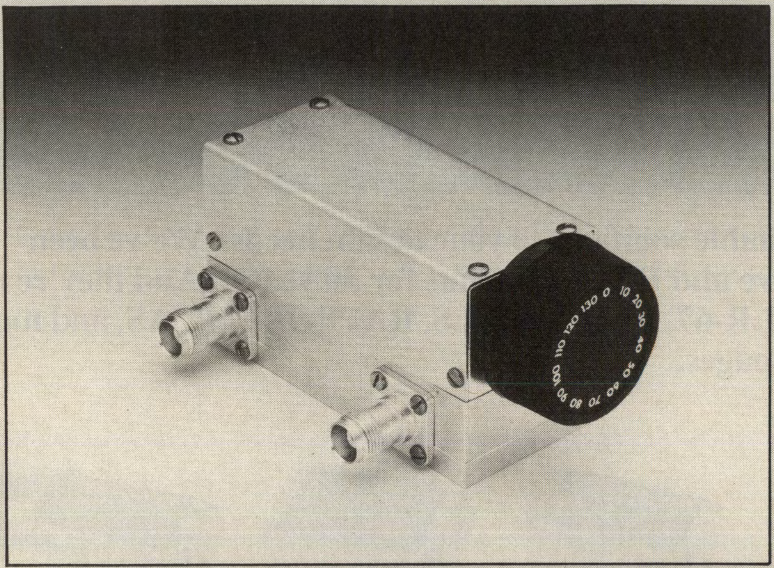

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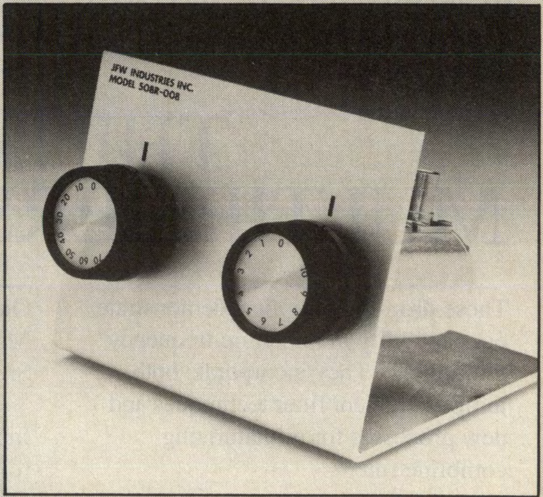

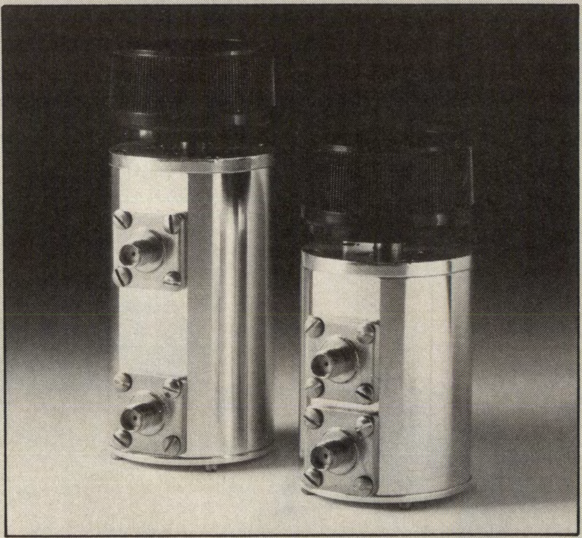

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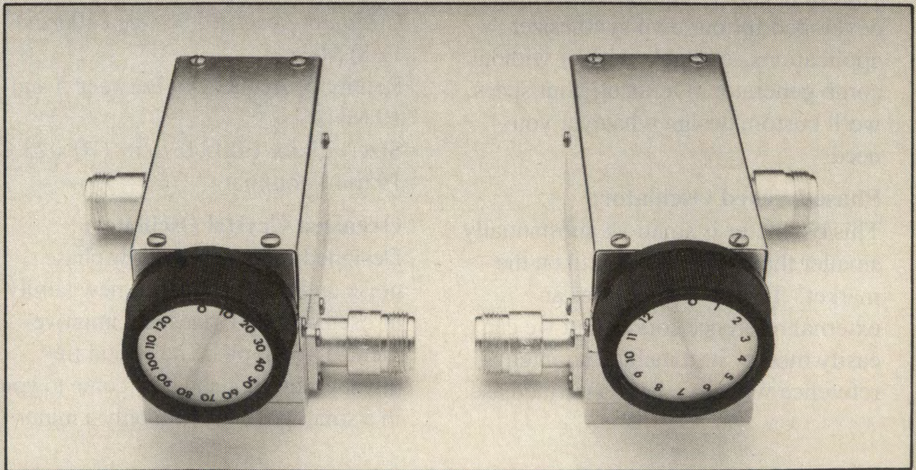

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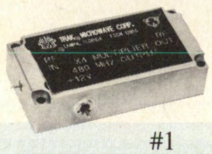
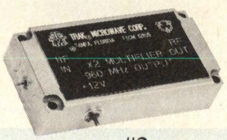
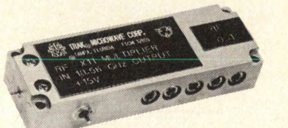

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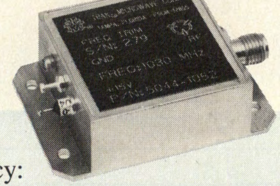


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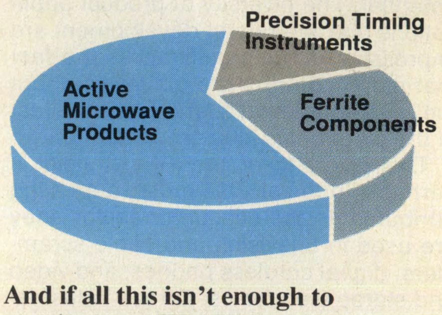

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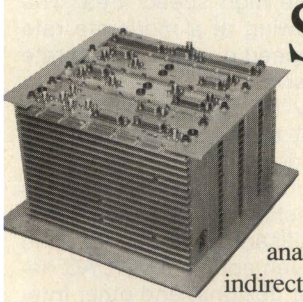

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INFO/CARD 27

A Bright Future for RF ICs

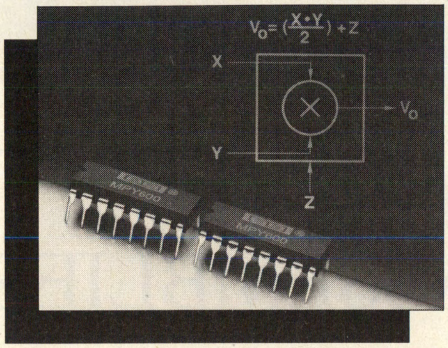
By Liane Pomfret and Charles Howshar
Assistant Editors

After discussions with several manufacturers of integrated circuits, it has become apparent that the RF integrated circuit market has tremendous potential. The diversity of product applications and the rate of development are impressive. More significant is the fact that manufacturers are optimistic about the RF IC market at a time when the rest of the electronics industry is stagnant.

The diversity of the RF IC field is perhaps one of its most appealing attributes. In the consumer sector, they are used in television and cable scramblers, digital cordless phones, and video and stereo electronics. On the industrial side their uses are more widespread: wire modem/wireless LANs, automatic test equipment, magnetic resonance imaging, test instrumentation, medical instrumentation, image scanners, computer peripherals and radar systems. New fields are constantly emerging, which makes this an ideal market for companies to expand into. William Gross, Design Manager for Linear Technology notes, "We're just getting into the high speed amplifier market, and project excellent growth in that market." Other manufacturers are looking down the road to HDTV. Aki Kaniel, Strategic Marketing Manager for Elantec comments, "We are developing chip sets now for HDTV which we think will certainly be a growing market, certainly not right now, but in two or three years."

An area of interest right now is analog to digital and digital to analog converters. Specifically, converters in the 12-bit and above range are in great demand. Alan Hansford, Product Marketing Manager for the Data Conversion Group at Comlinear hints that, "Converter speeds in the 12-bit range are easily going to double in the very near future." In addition, he believes that, "There will be heavy price erosion in the-12 bit product area much like the 8-bit devices three years ago."

One aspect customers are pushing for is higher performance — higher speed, power, precision, spectrum usage, and frequency in all areas of IC usage. Peter Predella, Technical Publicity Associate for Analog Devices, notes that, "The trend is towards higher resolution devices and also towards being able to use more of the RF



spectrum." Tom Munson, an Application Engineer for the RF/Linear/Radar Division at Plessey agrees, "Radar people want IFs to go to higher frequencies,... Front end people all want 900 MHz spread spectrum." According to Steve Pratt, Business Manager at Maxim Integrated Products, his company has responded to these demands, "Maxim's complementary bipolar process allows us to develop high speed, low power op amps and comparators which were previously unobtainable."

While there appears to be a move to "go digital," feelings are mixed as to the eventual direction of the market. Neil Albaugh, Manager of Advanced Products Development, the Military Products Division at Burr-Brown states, "We think the progress of the industry in moving towards a digital approach is very exciting." Alan Hansford sees future technology more in the analog area saying, "We find that the analog technology is going to be the big technology." One example he gives is for the concept of direct IF sampling, "Once a 12-bit ADC becomes twice as fast, direct IF sampling becomes marginally capable." This would be a boon to communications groups because of the reduction in the amount of interference and noise.

To become more competitive in the marketplace, some manufacturers are looking at their fabrication facilities or outside facilities for ways to improve their processes. Peter Predella states, "Right now some of our converter lines are being manufactured on an AT&T process. It's very advanced. We're looking at their process and trying to improve on it." The fabrication process is not the only area where companies are looking to improve. Aki Kaniel comments, "We recently installed a complete CAD system upgrade. Sun workstations, and an analog workbench. We are trying to perfect every part of the company." For others, it is a matter of changing or introducing a product line. William Gross notes "We just ended our fiscal year and and are looking forward to entering the high speed amplifier market." Dennis Monticelli, Design Manager for National Semiconductor, indicated that they are looking to move into the Japanese market, specifically in the computer peripherals area.

In addition, several companies are working on development of new processes. John Krehbiel, Manager of Business Development for Analog Signal Processing at Harris Semiconductor notes, "We're working on a new process which is still in development stage, that will be the fastest complementary bipolar process." Companies such as Linear Technology, National Semiconductor, Elantec, Maxim, and Burr-Brown all mentioned development of high speed complementary bipolar processes; some with higher voltages and others with faster speeds.

The general sentiment among IC manufacturers is one of optimism. According to Dennis Monticelli, "Our sales are definitely going up, there's a good market. But also National's entry into the market has helped." Aki Kaniel agrees, "The military market is flat, however the commercial market is growing." Peter Predella follows the same line of thinking, "At this point in time we're seeing RF as being one of the upticks in our sales. The entire industry is flat." Steve Pratt comments, "Our sales are going very well in the high speed area. The industry is growing at a moderate rate and our high speed product growth rate is very aggressive."

While many RF markets are struggling to adjust to the vacuum created by recent cutbacks in military spending, the RF integrated circuit market is growing. IC Technology is on the verge of some significant performance and process improvements in ADC/DAC designs, amplifier designs, and other integrated circuits. Given the prospect of some tough times for the electronics industry, it is good to see that there are a few bright spots in the industry. ***RF***

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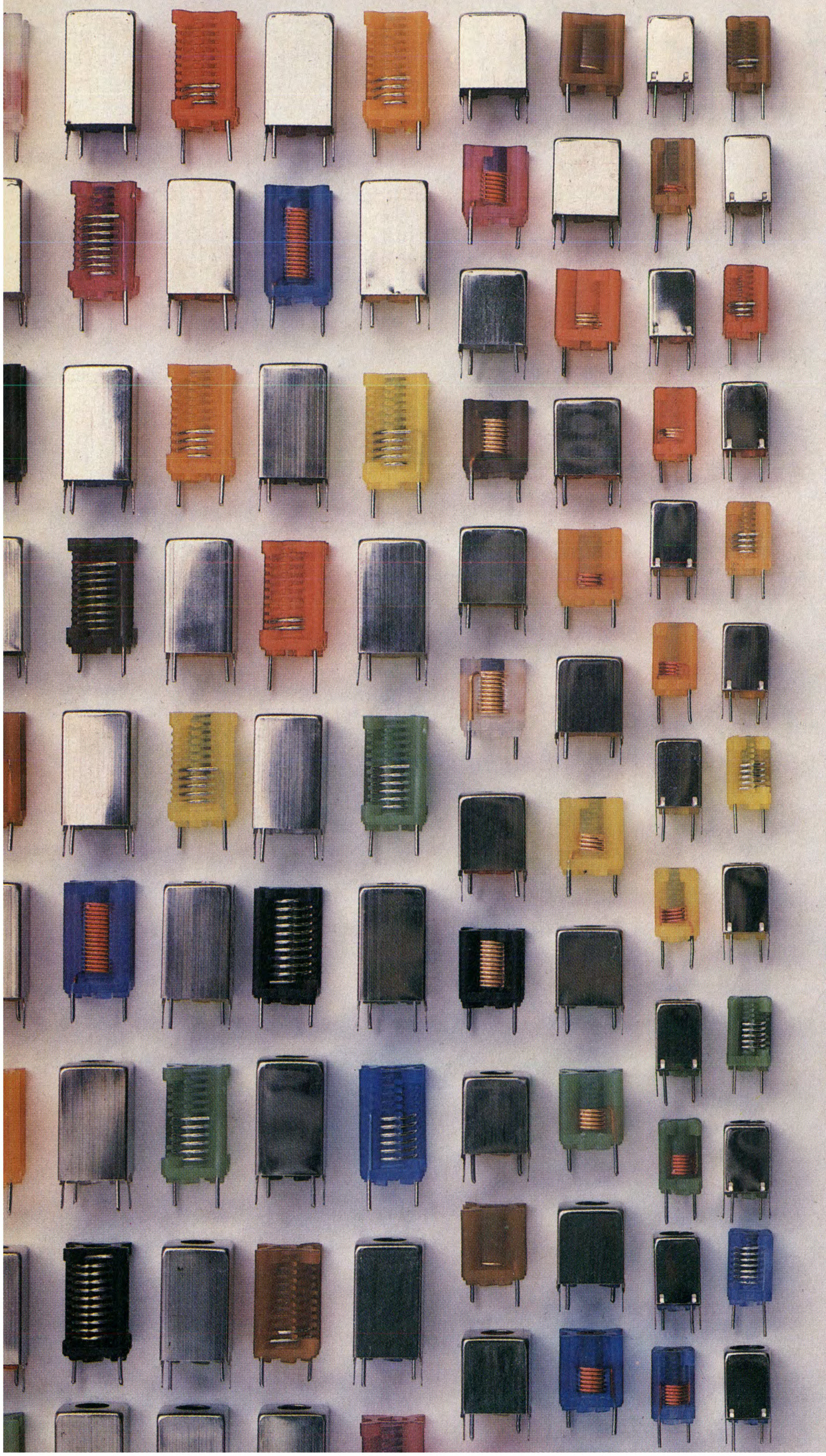

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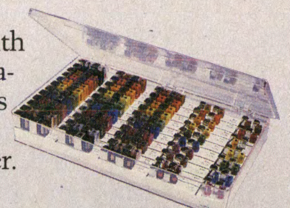

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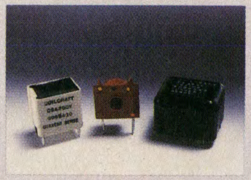

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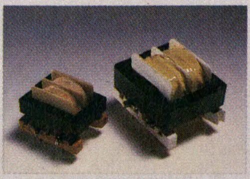

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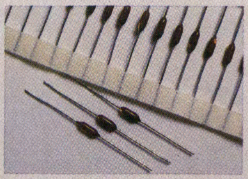

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INFO/CARD 29

Negative Gain-Single Pole Oscillators

By Leonard L. Kleinberg
NASA Goddard Space Flight Center

There is a class of oscillators that may be characterized as having a negative gain and a single pole response (NG-SP). The previously reported gate oscillators (1) are members of this class. These oscillators, while suitable for some applications, require considerable supply current for the higher frequencies. The purpose of this article is to continue the discussion on the technique of analyzing oscillator circuits and to describe a very simple NG-SP bipolar transistor oscillator that requires a small (2.5 mA) supply current. Also included, for purposes of demonstration, and perhaps utility, are operational amplifier oscillators.

An elementary model of a NG-SP system is depicted in Figure 1. As discussed in the August, 1989 edition of *RF Design*, the driving point admittance (DPA) of such a system is given by:

$$\mathrm{DPA} = \left[1 - \left(-\frac{kA}{s}\right)\right] Y_f = \left(1 + \frac{kA}{s}\right) Y_f \tag{1}$$

where $s = j\omega$ and A is in radians/sec. If Y_f is equal to $1/j\omega L$, the DPA becomes:

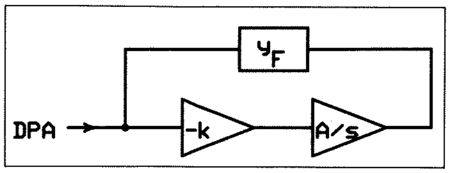

Figure 1. Simple model of a negative gain-single pole oscillator.

$$\mathrm{DPA} = \left(1 + \frac{kA}{j\omega}\right)\frac{1}{j\omega L} = \frac{1}{j\omega L} - \frac{kA}{\omega^2 L} \tag{2}$$

The DPA consists of two admittances, an inductive admittance and a negative conductance. In order to form an oscillator a capacitive admittance and a positive conductance must be added to the DPA, as shown in Figure 2.

The equations of the oscillation are:

$$j\omega C + \frac{1}{j\omega L} = 0 \quad ; \quad \omega^2 = \frac{1}{LC} \tag{3}$$

$$Y_{in} - \frac{kA}{\omega^2 L} = 0 \quad ; \quad Y_{in} = kAC \tag{4}$$

Notice that if C is chosen to set the frequency, where L might be the inductance of a crystal, Y_{in} will have to be adjusted to maintain the equality expressed in equation 4. This will be done for the bipolar transistor implementation of the fundamental system.

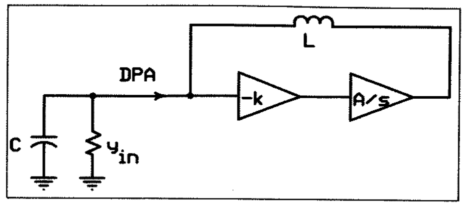

Figure 2. Driving point admittance (DPA) oscillator.

Figure 3a shows a possible, but improbable, embodiment of a NG-SP oscillator employing an operational amplifier as an integrator. Figures 3b and 3c demonstrate an improper and a proper configuration for a feedback analysis approach to the oscillator design. For the circuit of Figure 3a, the total admittance at point A is:

$$Y_t = j\omega C_1 + \frac{1}{R} + \mathrm{DPA} \tag{5}$$

$$= j\omega C_1 + \frac{1}{R} + \left(1 + \frac{1}{j\omega CR}\right)\frac{1}{j\omega L_1}$$

$$Y_t = j\omega C_1 + \frac{1}{j\omega L_1} + \frac{1}{R} - \frac{1}{\omega^2 L_1 CR}$$

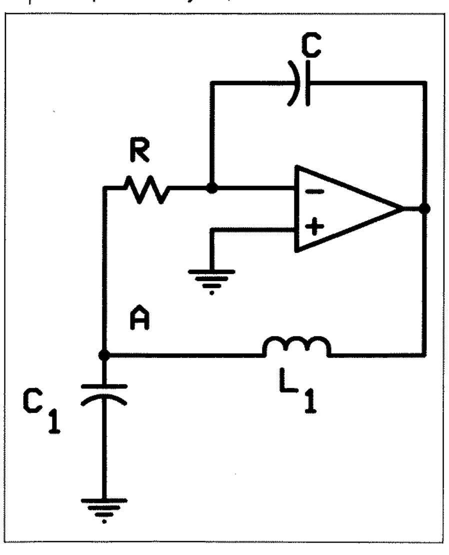

Figure 3a. Possible but improbable op amp NG-SP oscillator.

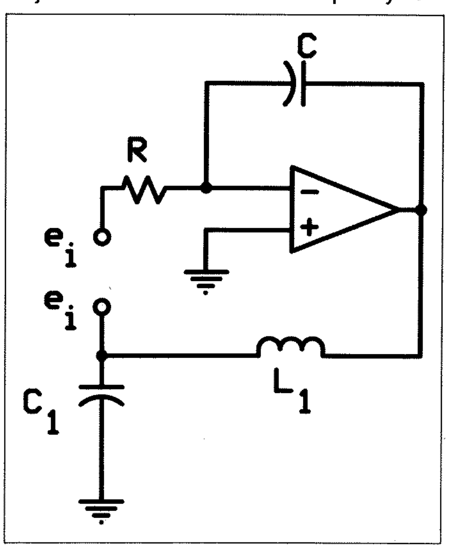

Figure 3b. Improper configuration for feedback analysis approach to oscillator design.

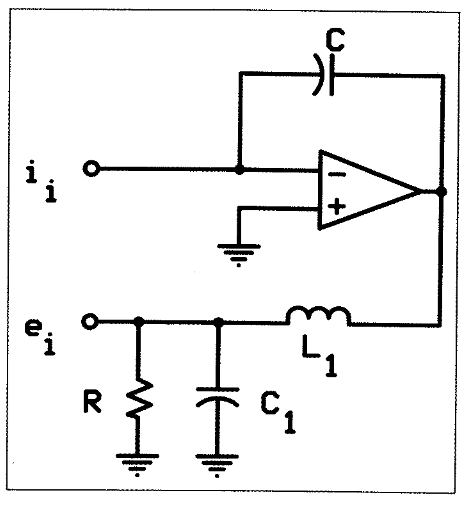

Figure 3c. Proper configuration for feedback analysis approach to oscillator design.

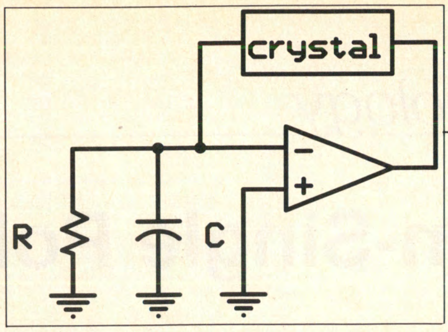

Figure 4a. Practical op amp NG-SP oscillator.

where A = 1/RC and k = 1. The conditions of oscillation are:

$$j\omega C_1 + \frac{1}{j\omega L_1} = 0 \quad ; \quad \omega^2 = \frac{1}{L_1 C_1} \qquad (6)$$

$$\frac{1}{R} - \frac{1}{\omega^2 CRL_1} = 0 \quad ; \quad C_1 = C \qquad (7)$$

The circuit of Figure 3b is an example of an incorrect method for the feedback analysis approach. The input signal e_i is amplified, and then attenuated by the feedback network to yield the same input e_i. The expression for this approach is:

$$e_i\left(-\frac{1}{j\omega CR}\right)\left(\frac{\frac{1}{j\omega C_1}}{\frac{1}{j\omega C_1} + j\omega L_1}\right) \neq e_i \qquad (8)$$

The circuit of Figure 3c is correct for analysis as C_1 is paralleled by R, and the input to the amplifier is a current.

$$i_i\left(-\frac{1}{j\omega C}\right) = \frac{-e_i}{R}\left(\frac{\frac{1}{j\omega L_1}}{\frac{1}{j\omega L_1} + j\omega C_1 + \frac{1}{R}}\right) = e_i \qquad (9)$$

This expression has the same solution as equations 6 and 7. The DPA avoids problems of this nature. Since an operational amplifier is an example of a NG-SP device, the oscillators of Figure 4 are practical circuits. Notice that the circuit of Figure 4b, which is similar to

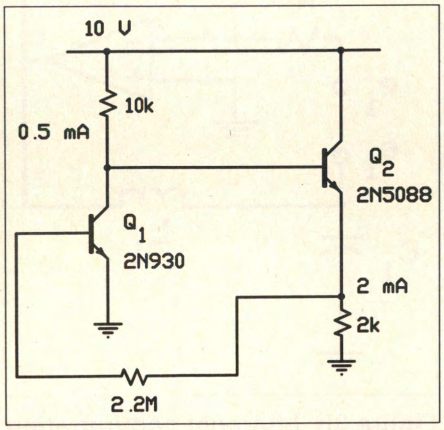

Figure 5. Approximation of a NG-SP oscillator.

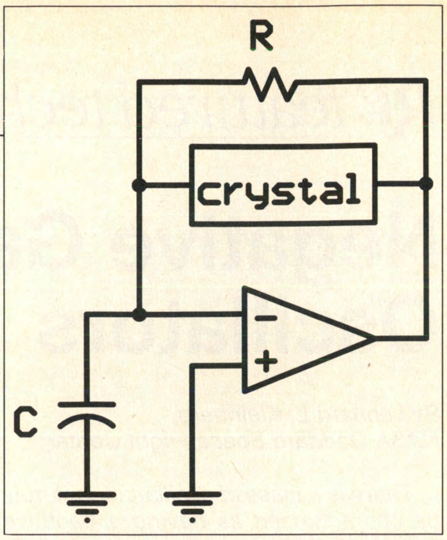

Figure 4b. Practical op amp NG-SP oscillator.

the gate oscillator in Reference 1, is capable of overtone operation. The total admittance at the inverting terminal of the op amp in Figure 4a is:

$$\begin{aligned} Y_t &= j\omega C + \frac{1}{R} + DPA \\ &= j\omega C + \frac{1}{R} + \left(1 - \frac{j\omega_G}{\omega}\right)\frac{1}{j\omega L_x} \\ &= j\omega C + \frac{1}{j\omega L_x} + \frac{1}{R} - \frac{\omega_G}{\omega^2 L_x} \end{aligned} \qquad (10)$$

where $k = j\omega_G/\omega$, F_G = GBW product, and L_x = crystal inductance. The equations of oscillation are:

$$j\omega C + \frac{1}{j\omega L_x} = 0 \quad ; \quad L_x = \frac{1}{\omega^2 C} \qquad (11)$$

One should recognize that C is the required loading capacitance of the crystal.

$$\frac{1}{R} - \frac{\omega_G}{\omega^2 L_x} = 0 \quad ; \quad \frac{1}{R} = \omega_G C \qquad (12)$$

For a given C, R is adjusted for optimum performance. For the circuit of Figure 4b the total admittance at the inverting terminal is:

$$\begin{aligned} Y_t &= j\omega C + DPA \\ &= j\omega C + \left(1 - \frac{j\omega_G}{\omega}\right)\left(\frac{1}{j\omega L_x} + \frac{1}{R}\right) \\ Y_t &= j\omega C + \frac{1}{j\omega L_x} - \frac{j\omega_G}{\omega R} + \frac{1}{R} - \frac{\omega_G}{\omega^2 L_x} \end{aligned} \qquad (13)$$

The equations of oscillation are:

$$j\omega C + \frac{1}{j\omega L_x} + \frac{\omega_G}{j\omega R} = 0 \qquad (14)$$

$$\frac{1}{R} - \frac{\omega_G}{\omega^2 L_x} = 0 \qquad (15)$$

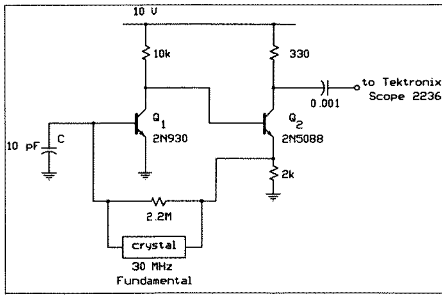

Figure 6a. Easiest possible approximation configuration.

This is somewhat different in that an additional reactive is present. As in the case of the gate oscillator, this configuration will permit overtone operation by properly selecting R.

For these oscillators, the slew rate must be compatible with the frequency of oscillation. A stable, low frequency oscillator, 250 kHz to 2 MHz, is obtained by using the Motorola MC34001, having a slew rate of 13 V/μs and an F_G of 4 MHz. For a higher frequency oscillator, Analog Devices AD847 is suitable, having a 300 V/μs slew rate and an F_G of 50 MHz.

The circuit of Figure 5 approximates a NG-SP oscillator. The voltage gain bandwidth product may be represented by:

$$F_G \cong \frac{20\ V\ F_t}{B} \tag{16}$$

where F_t is the transitional frequency of the bipolar transistor (Q_1) and B is the low frequency β. V is the supply voltage and the collector voltage is V/2. The 2N930 has an F_t of 30 MHz at 0.5 mA collector current and a β of 150, yielding an F_G of 40 MHz. The supply voltage V is 10 volts. The 2N5088 provides a low output impedance over a wide bandwidth (400 MHz). The system takes 2.5 mA at 10 volts.

The circuit of Figure 6a, the easiest possible configuration, did not perform well. The input admittance of Q_1 does not resemble that of an op amp or a CMOS inverter. The input admittance is primarily capacitance and must be considered when ordering the crystal. An input conductance is still required.

The circuit of Figure 6b includes a resistor R_3 across the crystal to provide suitable conductance at the input terminal and a favorable inductive admittance. Refer to equation 13. This was adjusted until an optimum value was obtained. The stability improved to where the short term stability was one part in 10^8 (1 sec). The circuit of Figure 6b was also used with a third overtone crystal at 35 MHz. The value of R_3 was 1.5 kohms. A larger value than 1.5

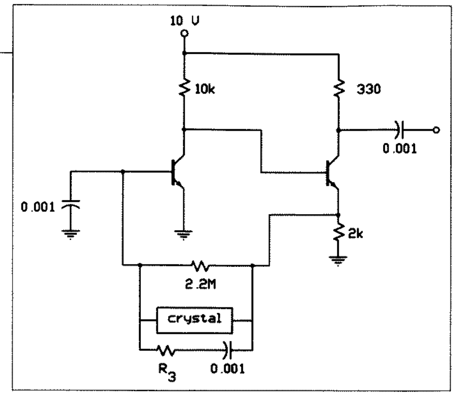

Figure 6b. Much better approximation of a NG-SP oscillator.

kohms produced oscillations at the fundamental frequency — 11.7 MHz. The mechanism by which the overtone was extracted is explained in (1) and is derived from equation 13. The inductive admittance, $-j\omega_G/\omega R$, dominates the parallel combination of itself with $j\omega C$ at the fundamental frequency, while at the third overtone the converse is true, and oscillations may take place.

Summary

Negative gain-single pole configurations are useful in synthesizing a variety of oscillators that are easily designed and may be classified as medium precision circuits. Previously, CMOS inverters were considered as the active device, but now operational amplifiers and bipolar transistors have been included. The techniques of analyzing oscillators by the driving point admittance method has been employed, and is the preferred technique in that anomalies may be avoided. Various embellishments of the circuits are possible, and one need not be confined to the circuits discussed. These oscillators were operated near the F_G of the system and experience has shown the performance to be better at about an octave below that. By re-biasing the transistor so that 1 mA flows through Q_1 (5kohm load resistor) the oscillator at 30 MHz was more stable, as was the 20 MHz oscillator at the same operation conditions. *RF*

References

1. Leonard Kleinberg, "An Analysis of Inverter Crystal Oscillators," *RF Design*, August, 1989, pp. 28-32.

About the Author

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Comments on MIL-STD-2000

By Robert Mouck and
Stanley Kaat
Bliley Electric Company

MIL-STD-2000 has replaced DOD-STD-2000, WS6536, and several earlier attempts to upgrade and control soldering and workmanship in circuit assemblies. It is rapidly becoming the standard for high-reliability assemblies in both "military" and "civilian" applications. We have demonstrated that a small manufacturer can comply with MIL-STD-2000, but compliance takes time for preparations and can add significantly to costs. We are presenting some comments on our experiences with it. While we are not presuming to critique MIL-STD-2000, we recommend that it be used only where it is actually needed.

MIL-STD-2000, "General Requirements for Soldered Electrical and Electronic Assemblies," consists of six sections. The Baseline portion defines the general requirements; the five Tasks prescribe the procedures which are to be used. Task A = Training, B = Design, C = General Processes, D = Hand Soldering, and E = Machine Soldering. Two additional Tasks, F and G are combinations of the others; F = Everything but Design and G = Everything.

Baseline Requirements

The "Baseline requirements" of MIL-STD-2000 are not significantly different from the old high reliability specifications MIL-S-45743 and MIL-S-46844, and are very similar to WS6536 and DOD-STD-2000. Even at the Baseline level, MIL-STD-2000 is a very comprehensive document. Proving compliance with its requirements can demand more preparation, more monitoring, more written procedures, and more record keeping than is apparent from a casual reading of the MIL-STD-2000 document. Do not confuse this with "high-quality commercial" work; the Baseline is a definite program in itself. The various Tasks add complexity, but the program remains the same.

Tasks

Task A requires an on-staff Category C expert be trained and certified at one of a few special military schools. In our case, our applicant waited seven months from acceptance until actual enrollment. Upon graduation, the "Cat C" conducts training for operators and inspectors, using the government supplied 400 page training manual. Tasks B, C, D, and E define the design and process requirements. In effect, they provide the details of how the general requirements are to be achieved; in only a few instances do they add to or tighten the actual requirements. But, of course, they do demand tighter controls, more documents, more tests, and more records.

Application

MIL-STD-2000 is intended to be applied in either of two versions, the Baseline requirements portion only, or the Full program (which itself has two variations, differing only in whether or not the Design Task is included.) The Baseline is designed to be invoked simply by reference to MIL-STD-2000. The Full version is to be invoked as either Task F or as Task G. Perhaps the Baseline might be considered the "civilian" version and the Full program, the "military" version. Both intend to achieve almost the same goals, but the "military" version prescribes procedures as well as results.

Preparations

Much of what must be done to prepare for compliance with MIL-STD-2000 depends, in our experience, on two factors. First, the experience base of the organization and its people is critical, because fitting MIL-STD-2000 into a group already accustomed to military requirements and procedures is difficult enough; fitting it into a group in which all the supplemental activities had to be developed simultaneously would be a monumental undertaking. Second, much of what must be done to demonstrate compliance with MIL-STD-2000 is still subjective (even where requirements are clear there can be differences of opinion as to what constitutes adequate demonstrations of compliance.) A thorough mutual understanding and agreement with the administering official needs to be made early in the project.

Our Experience

In our case, we already had in place a fully-documented quality system and we were accustomed to doing similar work. Even so, it took us a full year to complete our preparations. This was spent in determining the requirements, reaching agreement on what would be done to demonstrate compliance with each, preparing documents, acquiring equipment, and training people. We believe this thorough preparatory phase is crucial to success. We recommend that everyone who plans to install MIL-STD-2000 make adequate provisions for this preparation.

Documents

The MIL-STD-2000 program demands many documents and many records: operator certifications (including visual acuity and proficiency), detailed work instructions, equipment calibration, a documented ESDS program, environmental controls (temperature, humidity, illuminations, cleanliness), and ongoing records demonstrating compliance. Production lot documents must be comprehensive and must correlate to all other ongoing records. Even when these are incorporated into an existing system, this can be formidable.

Equipment

In addition to the facilities obviously necessary to accomplish the work, MIL-STD-2000 can demand some special items. For us, this included a certified photometer so light levels at work stations could be monitored and recorded, similarly to the ongoing monitoring and recording of ambient temperature and humidity. A certified stopwatch was also needed. Certification, of course, demands traceability to NIST (NBS). As with the documents, calibration and certification are less burdensome if they can be included in an existing system.

Effect on Costs

In addition to the increased overhead, adherence to the MIL-STD-2000 program inevitably adds to the direct cost of processing. Our experience indicates that, initially at least, per-unit labor costs are 50 percent higher, mostly because of the detailed inspections and associated rework. There is a corresponding penalty in the time needed to process a lot. Imposition of MIL-STD-2000 will obviously significantly increase costs.

Recommendations

We now know that a small manufacturer can comply with MIL-STD-2000. As manufacturers, we have a plea for you designers, the familiar request, "Don't over specify!" Don't include any requirements you don't actually need. That's always good advice, but particularly so for MIL-STD-2000 . And please don't slip in the MIL-STD-2000 requirements indirectly, as by citing MIL-STD-454 Requirement 5. If you really need MIL-STD-2000, we small manufacturers can give it to you. But you will have to tell us in time so we can prepare, and you'll increase your costs and ours. *RF*

About the Authors

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Loading Conditions for High Speed Logic Systems

By Alfred Dantas
Valpey-Fisher Corporation

Logic devices are now being developed to operate at microwave frequencies exceeding 2.5 GHz. In order to utilize high speed logic devices successfully, careful system timing design and good circuit board layout are necessary, especially when using multilayer boards. Pitfalls and problems not encountered when using logic devices at low frequencies will be discussed here. This information will be valuable to system designers using clock oscillators in high speed logic systems (clock frequency typically > 25.000 MHz). Applicable clock oscillator technologies include TTL, HCMOS, ACMOS, ECL (10K and 100K devices).

A clock oscillator used in a logic system will be connected to the system either by short lines on a printed circuit board, coaxial lines, or longer lines on a back plane. As the clock frequency is increased, the interconnected media must be treated as transmission lines with a characteristic impedance and a propagation constant. The transmission lines store energy of a magnitude determined by the source impedance. The stored energy must be dissipated by a terminating device in order to reduce reflections. The oscillator can be considered an RF source, and the effects of improper line termination on the clock oscillator will therefore be discussed in terms of transmission line concepts.

When a source is mismatched with a load, a reflected voltage will arrive back at the source, or generator. The reflection coefficient at the source will determine the response of the generator to the reflected voltage, V_r. The reflection coefficient at the source is determined by the characteristic impedance of the transmission line, Z_0, and the source resistance, R_s. The reflection coefficient is given by the equation:

$$\varrho_s = \frac{R_s - Z_0}{R_s + Z_0} \qquad (1)$$

A reflected voltage arriving at the source will not be reflected back to the load if the source impedance matches the line impedance. Multiple reflections will occur if neither the source impedance nor the terminating impedance match the transmission line impedance, Z_0. The reflected voltage at each end of the line approaches a steady state value with each succeeding reflection. If, however, the source reflection coefficient and the load reflection coefficient are of opposite polarity, the reflections alternate in polarity causing the signal voltage to oscillate about a steady state value. This effect is often referred to as "ringing". A distorted waveform will be observed if the signal rise time is long compared to the transmission line's delay (Figures 12a,12b). There are five types of transmission lines and driving conditions for each of the five types are discussed here.

Short lengths of unterminated transmission lines may be driven successfully. The length is defined by the expression:

$$1 = \frac{T_r}{2T_{pd}} \qquad (2)$$

where T_r is the rise time (ns) and T_{pd} is the propagation delay (ns/in).

The maximum current the active device has to deliver to the next circuit can also be obtained in terms of the above expression. For an example where T_r = 3 ns, the maximum length is 10 inches for a trace on a G–10 epoxy glass printed circuit board. A voltage wave propagating down the line will have the maximum amplitude which the source can drive into Z_0 (Figure 1). Reflections may not present a serious problem if the source current, I_s, is less than I_{max}. Reflections will occur if the line is longer than the value defined by I_{max}. The reflections will drive the input and output signal voltages below the ground voltage level.

Series terminated lines have limited uses due to voltage drops in the series resistor values, R_T of Figure 2, in the low voltage state. The resistor also reduces the noise margin in receivers.

For parallel terminated lines there are four line terminating configurations that may be used successfully:

Z_o to V_{cc}. This termination configuration will draw current from V_{cc} when the output is Low (Figure 3).

Z_o to Ground. The termination in this configuration will draw current when the output is High (Figure 4).

Thevenin Equivalent Termination. This type of termination will draw current from the RF source and from V_{cc} when the output is in either the High or Low state (Figure 5). This configuration will also reduce the noise margin. The

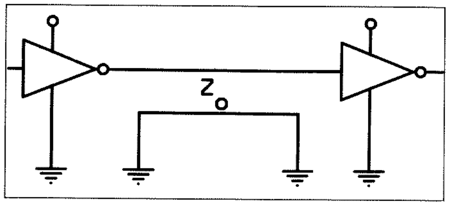

Figure 1. An unterminated transmission line.

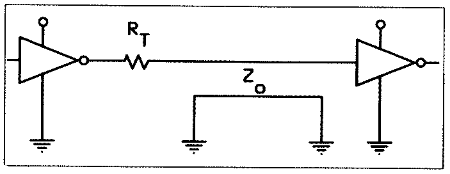

Figure 2. Series terminated transmission line.Typical R_T values of 10 to 50 ohms may be used.

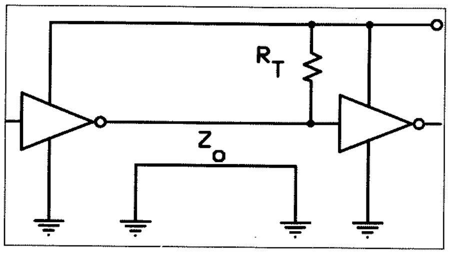

Figure 3. Z_o toV_{cc}.

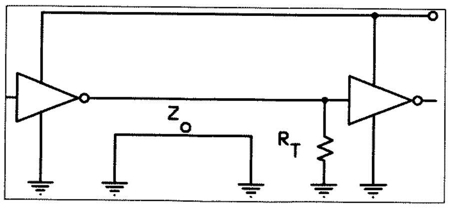

Figure 4. Z_o to ground.

resistors, R_{T1} and, R_{T2} will set the quiescent operating line voltage and the Thevenin equivalent terminating impedance. This configuration is used when only one supply voltage is available. It can be used successfully with Tri-state logic.

AC Termination. This configuration uses a capacitor in series with the terminating resistor to block DC current (Figure 6). The load is therefore defined as $Z_o = R + X_c$. The reactive impedance of X_c must be small compared with Z_o at the operating frequency for the load to appear predominantly resistive. This termination does not draw any DC current with the output in either state. The quiescent line voltage can be established at V_{cc} or GND by a large value "pull up" or "pull down" resistor applied to the appropriate supply rail.

Transmission Line Media

All logic devices transmit information along a conductive medium. This medium must be considered as a transmission line when used with all high speed logic systems. The characteristics of the transmission line can therefore be defined in terms of the conductor and dielectric parameters of the transmission medium. Parameters critical to proper calculation of transmission line characteristics are:

h = dielectric thickness
t = thickness of conductor trace
l = length of conductor trace
K = dielectric thickness between ground planes
w = width of conductor trace
ε_r = dielectric constant

The characteristic impedance, Z_o, of some of the more commonly used printed circuit configurations used in the computer industry will be outlined here. Variations of the equations used to determine the characteristic impedance, Z_o, of the various types of transmission lines have been discussed in greater detail by other authors (4).

The characteristic impedance of the micro-strip line can be obtained from Figure 7 and the following equations.

$$Z_0 = \frac{87}{\sqrt{\varepsilon_r + 1.41}} \ln\left(\frac{5.98h}{0.8w + t}\right) \text{ohms} \quad (3)$$

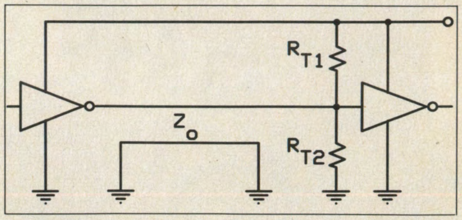

Figure 5. Thevenin equivalent termination.

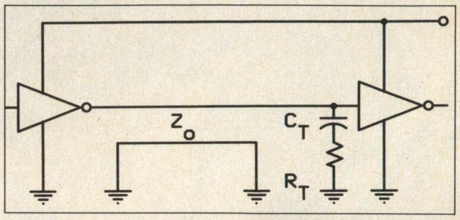

Figure 6. AC termination.

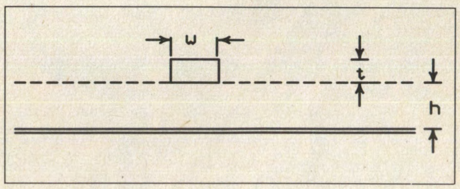

Figure 7. Micro strip-line.

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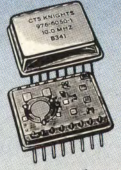

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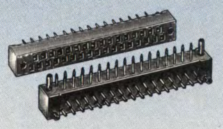

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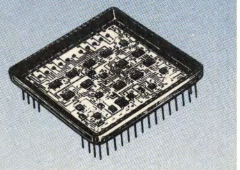

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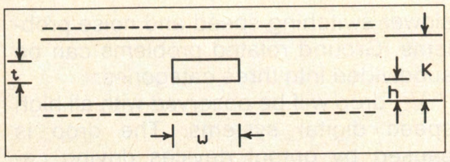

Figure 8. Stripline (tri-plate).

$$T_{pd} = 1.017\ \sqrt{0.474\varepsilon_r + 0.67}\ \text{ns/ft.} \tag{4}$$

The characteristic impedance of a stripline conductor can be obtained from Figure 8 and the following equations.

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}}\ \ln\left(\frac{4K}{0.67\pi w(0.8 + t/w)}\right)\ \text{ohms} \tag{5}$$

$$T_{pd} = 1.017\ \sqrt{\varepsilon_r}\ \text{ns/ft} \tag{6}$$

The characteristic impedance of two parallel transmission lines can be obtained from Figure 9 and Equations 7 and 8.

Characterizing the interaction of several lines running parallel to each other will be more difficult.

$$Z_0 = \frac{120}{\sqrt{\varepsilon_r}}\ \ln\left(\frac{\pi h}{w + t}\right)\ \text{ohms} \tag{7}$$

$$T_{pd} = 1.017\ \sqrt{0.475r + 0.67}\ \text{ns/ft.} \tag{8}$$

The characteristic impedance of coaxial cable depends upon the diameters of the conductor, the shield and the dielectric medium separating them. Z_o can be obtained from Figure 10 and equation 9.

$$Z_0 = \frac{60}{\sqrt{\varepsilon_r}}\ \ln\left(\frac{D}{d}\right)\ \text{ohms} \tag{9}$$

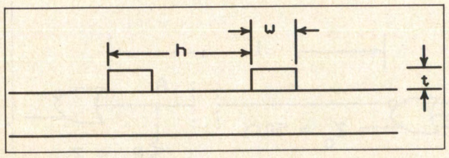

Figure 9. Side-by-side transmission lines.

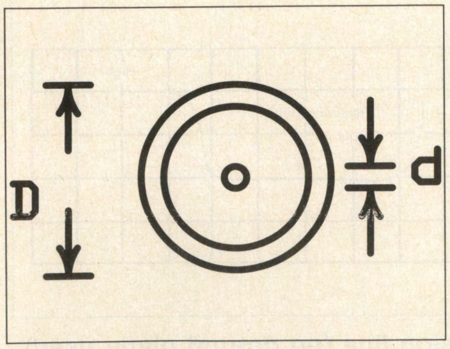

Figure 10. Coaxial cable.

And finally, for a twisted pair or ribbon cable, the characteristic impedance is derived from a more complex relationship (Figures 11a,11b). In the following equations each of the parameters R_o, L_o, G_o, and C_o are defined in terms of unit lengths.

$$Z_0 = \frac{120}{\sqrt{\varepsilon_r}}\ \ln\left(\frac{2D}{d}\right)\ \text{ohms} \tag{10}$$

$$Z_0 = \sqrt{\frac{R_0 + j\omega L_0}{G_0 + j\omega C_0}} \tag{11}$$

where R_o = Resistance in ohms per unit length,

L_o = Inductance in Henries per unit length,

G_o = Conductance in mhos per unit length,

C_o = Capacitance in Farads per unit length.

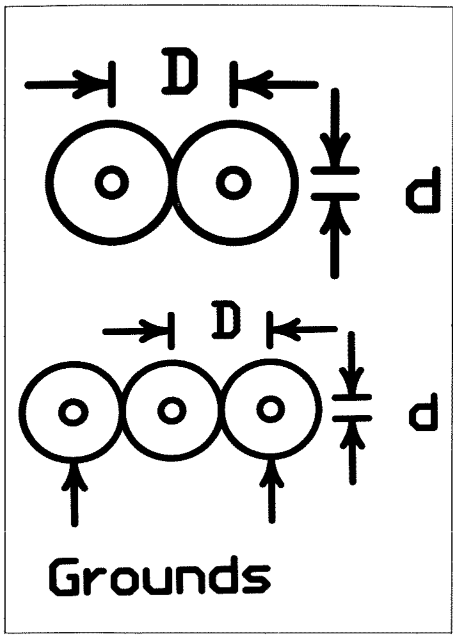

Figure 11. (a) Twisted pair. (b) Ribbon cable.

These expressions can be simplified to:

$$Z_0 = \sqrt{\frac{L_0}{C_0}} \quad (12)$$

$$T_{pd} = \sqrt{L_0 C_0} \quad (13)$$

assuming $G_o = R_o = 0$. Note that Z_o is real, not complex. The impedance is resistive and is not a function of length.

Loads and Terminations

Several more issues have to be taken into consideration when designing high speed logic systems. Some of the more important ones are:

Propagation delay in low and medium speed systems may be ignored, but becomes important in high speed systems. For instance, ECL logic operating at a speed of two nanoseconds introduces one equivalent "gate delay" into the system for every foot of interconnecting conductor trace.

Electrical noise generation, "crosstalk" between adjacent conductor traces, poor terminations and reflections are some of the agents which produce distorted waveforms that could lead to false triggering. Examples of a typical waveform of an unterminated transmission line (no ground used) and a properly terminated transmission line (ground plane used) are shown in Figures 12a,12b. The low impedance, emitter-follower outputs of ECL logic facilitate the use of transmission line practices but require the use of two DC power supply voltages. The power supply voltage levels for ECL logic are V_{tt}, V_{cc}, and V_{ee}. The voltage levels for the more commonly used negative logic are V_{cc} = 0 Volts (GND), V_{tt} = −2.0 Volts, V_{ee} = −5.2 Volts. The voltage levels for positive logic are V_{cc} = +5.2 Volts, V_{tt} = +3.2 Volts, and V_{ee} = 0 Volts (GND). These voltage levels facilitate the use of a 50 ohm (Z_o) transmission line system with a terminating load of 50 ohms. However, logic systems utilizing the transmission line concept but having characteristic impedances other than 50 ohms have been used successfully.

As logic speeds increase, the driven load appears to be more capacitive. Each additional load added to the transmission line causes a reflection with a polarity opposite to that of the incident wave. Reflections from adjacent loads would tend to overlap if the time required for the incident wave to travel from one load to the next is equal to or less than the signal rise time. An ECL oscillator would be capable of driving several devices provided appropriate RF matching techniques were used.

Design Considerations

The successful operation of high speed logic systems depends upon the correct implementation of RF techniques. The rise and fall times of high speed logic systems become a primary concern in printed circuit layout. Poor circuit layout techniques with insufficient ground planes could result in crosstalk, slower switching speed and noise problems. Ground related problems can be subdivided into three categories:

V_{cc} drop will be observed with all high speed digital systems. The drop is caused by output devices driving capacitive loads. The remedy for this problem is to provide more bypass capacitance to ground.

Coupling via ground paths adjacent to both signal source and loads, and crosstalk caused by signal paths can both be reduced by using correct RF printed circuit board techniques.

Most high speed logic systems are produced on double sided or multilayer printed circuit boards without regard to the physical placement of ground planes relative to signal lines. This means that signals or RF noise generated externally or by one of the circuits on the board could appear in other circuits on the board even though there is no apparent means of crosstalk between them. Decoupling each logic package is therefore necessary to reduce the effects of noise and crosstalk. Two forms of crosstalk may exist in a system, forward and reverse. The causes of both types are similar but their effects on the system are significantly different. These corrective measures may be used to reduce the effects of crosstalk:

- Maximize the spacing between signal lines.
- Minimize the spacing between the signal lines and ground.
- Use a ground strip between the cross-talker or cross-listener.

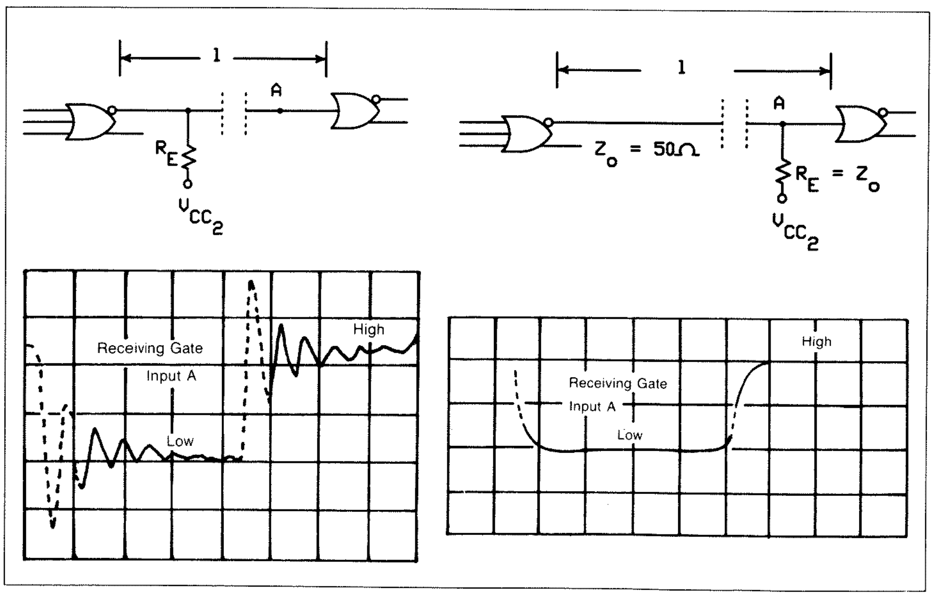

Figure 12. (a) Unterminated transmission line (no ground plane used). (b) Properly terminated transmission line (ground plane added.)

Ten Axioms for High Speed Circuit Designs

1. Minimize trace lengths of unterminated lines between logic gates.
2. Terminate the transmission lines with resistive terminations of the same value as the characteristic impedance of the lines.
3. Always terminate the transmission line at the input to another gate and not the input to the transmission line (Figures 3,4,5,6).
4. The terminating bias resistor for ECL oscillators should be placed at the input to the gate being driven by the oscillator.
5. Use buffer gates to isolate the ECL oscillator from the rest of the circuit.
6. RF power distribution techniques should be used with ECL oscillators driving more than one gate.
7. RF decoupling of the oscillator at the DC supply pin is essential for successful operation.
8. Decouple each logic package at the supply pin.
9. Ensure the characteristic impedance of the signal traces are maintained throughout the circuit.
10. Avoid discontinuities in the signal traces. A step transition in a signal trace represents a change of impedance at the discontinuity.

In conclusion, the successful operation of high speed logic circuits will be determined by careful design and layout of the circuit. Attention to RF circuit design will also be an important contributing factor. *RF*

References

1. Fairchild, F100K ECL User's Handbook, 1986.
2. Signetics, ECL 10K/100K Data Manual, 1986.
3. William R. Blood, Jr., *MECL System Design Handbook.* 1988.
4. T. C. Edwards, *Foundations for Microstrip Circuit Design,* John Wiley and Sons, 1981.
5. Paul L. Mathews, *Choosing and Using ECL,* R. R. Donnelley and Sons, 1984.
6. Fairchild, Fast Application Handbook, 1987.

About the Author

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Piezoelectric Devices Conference Marks its 12th Year

The newly renamed Piezoelectric (formerly Quartz) Devices Conference and Exhibition will be held September 25-27 at the Westin Crown Center in Kansas City. Sponsored by the Piezoelectric Subdivision of the Electronic Industries Association, the conference highlights manufacturing and engineering developments in crystal, SAW, and ceramic resonator oscillators, filters, and other circuits.

For the past seven years, proceeds from the conference and exhibition have been used to support research programs at Northern Illinois University, Oklahoma State University, and the University of Central Florida. The research includes work on numerous manufacturing, quality control, measurements, and performance aspects of piezoelectric devices. Results of some of these research projects are included in the technical papers.

The following papers are scheduled for this year's conference:

Tuesday, September 25
8:30 to 11:45 a.m.

- International Round Robin on Alpha Measurements of Cultured Quartz
- Effects of Etch Channel density on Chemically Shaped Very High Frequency Crystals
- Comparison of Nonlinear Materials Constants of Quartz From Different Methods of Measurement
- Influence of Quartz Material on the Photoetch Quartz Resonator
- Automated Detection and Identification of Crystal Defects With Digital Image Processing Techniques
- Surface Mount Miniature Quartz Crystal Resonators
- Thermal Hysteresis in Quartz Resonators - A Review

Tuesday, September 25
1:15 - 4:45 p.m.

- Generating Surface Figure and Finish on Quartz - Part 3
- Polishing to Flatness
- Updating Lapping Operations Through Improved Abrasive Controls
- Aging Considerations for the Ceramic Flat Pack Resonator

Wednesday, September 26
8:30 to 11:45 a.m.

- A Guided Tour of the EIA Guide to the Measurements of Equivalent Electrical Parameters of Piezoelectric Crystal Units
- Load Resonance Parameters of Crystals Without Use of a Load Capacitor
- New Network Analyzer Hardware and Software for EIA 512 Compatible Crystal Resonator Measurements
- Discussion of Measurement Problems and Possible Modifications to the Equivalent Circuit for Very High Frequency Resonators
- Orientational Dependence of "True" SC Cuts
- Method for the Reduction of Phase Noise in Oscillators
- Minimizing Frequency Pulling of Crystal Oscillators Due to Power Supply Variations

Wednesday, September 26
1:30 to 4:10 p.m.

- Design of Crystal Filters Using Coupled-Triple Resonators
- A Generalized Maximally Flat Approximation
- A Beveled Crystal Design Equation
- Crystal Notch Filter Derived from the Second Order Equalization Circuit
- Miniaturized Oven Oscillators

Thursday, September 27
9:00 to 10:15 a.m.

- Report of the Activities of Technical Committee 49 of the International Electrotechnical Commission
- MIL-STD-2000 and You
- Training Requirements for MIL-H-38534

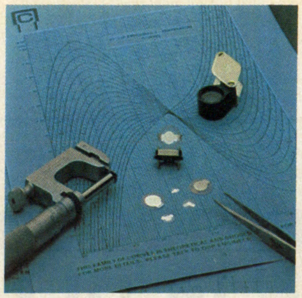
A meeting of the EIA P-11 Engineering Committee on Piezoelectric Devices is scheduled for Thursday afternoon, beginning at 1:00 p.m. The meeting is restricted to members and invited guests.

The exhibit hall will be open from 2:30 to 7:00 p.m. on both Wednesday and Thursday (September 25 and 26), and from 9:30 a.m. until noon on Thursday. Exhibitors (see sidebar) include crystal and oscillator manufacturers, plus tooling companies, raw material suppliers, and other firms involved in the manufacturing of crystals and related products.

This year's Kansas City gathering offers an opportunity for the quartz industry to catch up on technical developments and to mingle in a congenial atmosphere. Persons interested in attending the 12th Piezoelectric Devices Conference and Exhibition should see the announcement and registration form included elsewhere in this issue, or contact the EIA at (202) 457-4930, FAX: (202) 457-4985. Special airfare and hotel rates have been arranged for the conference. *RF*

Exhibiting Companies

A.T. Wall
Crystal Components
EG&G Frequency Products
Electro Dynamics Crystal
Emkay Manufacturing
Engineered Materials
Engineered Specialty Products
Fotofabrication Corp.
Gard Corp.
GSM Inc.
H.E.S. International
Highland Manufacturing Co.
Houston Electronics
K.C. Abrasive Co.
Kay, Inc.
M-tron Industries, Inc.
Marden Associates
P.R. Hoffman Machine Products
P.R. Hoffman Materials
Philips Components
Polaris Electronics Corp.
RF Design
Saunders and Assoc.
Sawyer Research Prod.
Schott Electronics
Shure Cut Crystal
Slee Superior
Thermo Dynamics
Transat Corporation
Universal Photonics
Valpey-Fisher Corp.
Voltronics Corp.

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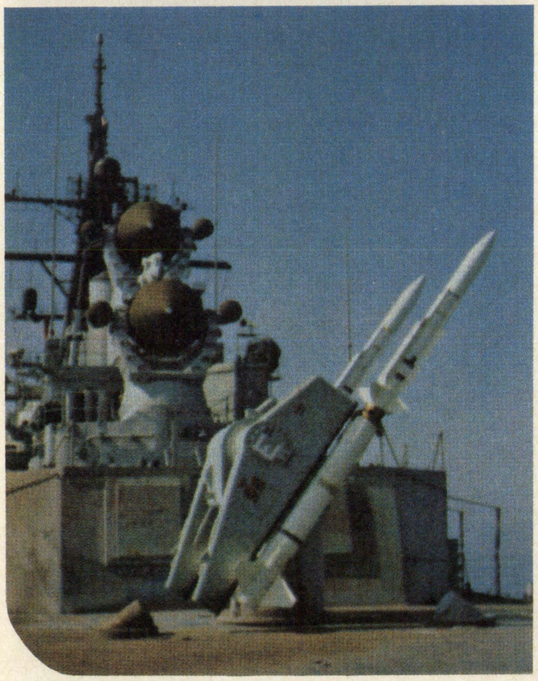
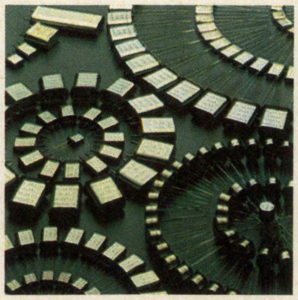
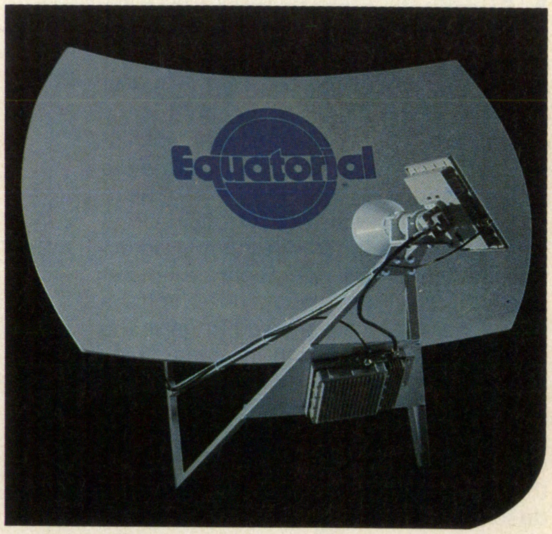





Equatorial

A Feedback Method for Reference Spur Reduction in PLLs

By John W. MacConnell and
Dr. Richard W.D. Booth

This paper describes a circuit that is extremely useful in reducing reference sidebands in phase locked loops or indirect frequency synthesizers which are caused by reference leakage from the active phase/frequency comparator. In excess of 40 dB reduction in reference spurious signals is routinely achieved without any impairment in loop dynamics due to spur reduction filters. The benefit of this circuit is that the necessary filtering required to obtain a given spurious signal level may be considerably reduced, especially in loops containing large division ratios such as microwave phase locked loops and microwave frequency synthesizers. In some cases, it has been impossible to achieve the desired spur level without the use of this circuit.

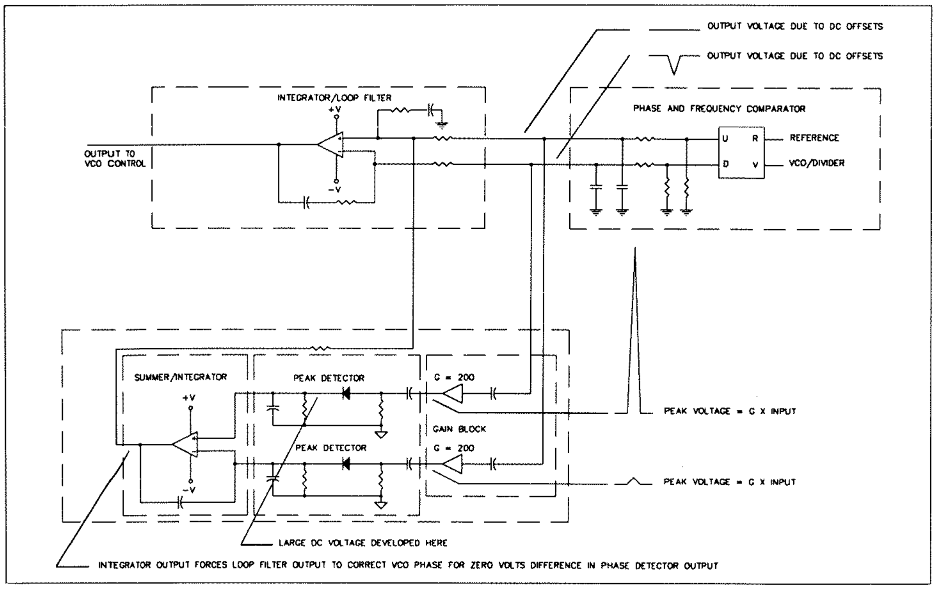


Figure 1. Block diagram.

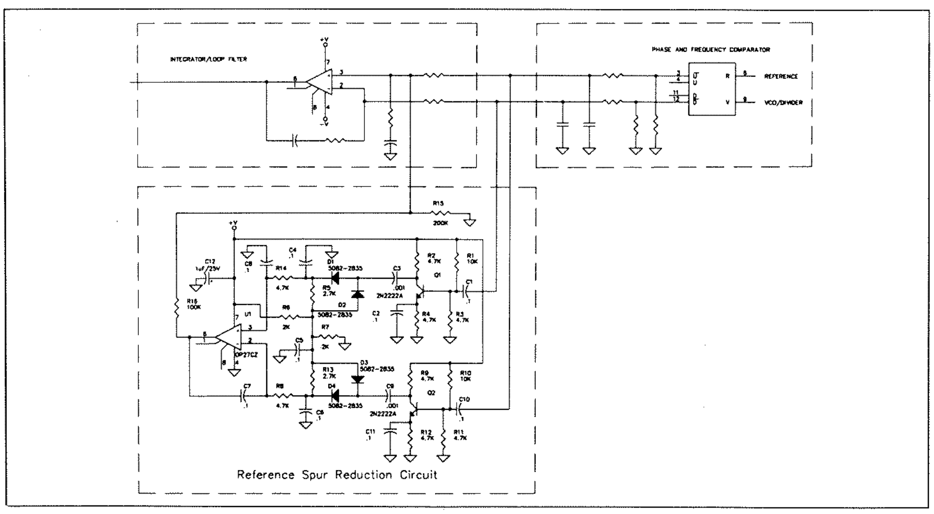


Figure 2. Schematic diagram of actual circuit.

One of the major sources of spurious signals in a phase locked loop (or indirect frequency synthesizer) is the reference leakage signal that appears at the output of the phase/frequency detector. For active phase/frequency comparators this leakage manifests itself as a narrow voltage spike at the Up or Down outputs (see Figure 1). These spikes are a result of DC offsets that exist within the phase/frequency detector itself, and the operational amplifier used in the loop filter. These voltage spikes generally are passed through the loop filter/integrator and in many cases are not attenuated much at all by the loop filter. The traditional method of dealing with these spurious signals is to use extensive filtering at the output of the loop filter/integrator. The circuit described in this paper does a great deal to remove these voltage spikes caused by DC offsets. It is simple and easily adapted to various loops and reference frequencies. It is not temperature sensitive and, because it is adaptive, does not require any adjustments.

Once a phase locked loop has reached a steady state condition, the differential input voltage to the operational amplifier used for the loop filter must be zero. If there were a net voltage at the input to the amplifier, its output would be ramping, up or down depending on the polarity of the input, indicating the loop has not yet stabilized. Therefore, if there is an offset, either within the phase detector, or within the amplifier, it must be nulled out. A normal phase locked loop achieves this by operating with a slight static phase offset within the phase detector, resulting in a voltage at its output which exactly nulls out the various offset voltages present in the circuit. Unfortunately, the way an active phase/frequency comparator achieves this is by putting pulses out on one or the other of its two outputs. The result is that one or the other of the phase

detector outputs generates narrow voltage spikes with an average DC value corresponding to the various DC offsets. Some amount of this pulse energy generally feeds through the loop filter and onto the VCO control line, causing reference spurs to be present in the output spectrum.

A few common techniques used to reduce reference spurs include:

1. Reducing the loop bandwidth.
   A. Undesirable: Increases loop settling time.
   B. Undesirable: Can increase phase noise.
2. Placement of lowpass and/or notch filters between the loop filter output and the VCO.
   A. Undesirable: Tends to cause loop instabilities.
   B. Undesirable: Increases circuit complexity.
   C. Undesirable: Notch filters often require adjustments at manufacture.

A New Approach

The new approach described here reduces the spurious signals by actively nulling out any offsets that occur.

Simply stated, this circuit measures the magnitude of the voltage spikes present at the output of the phase/frequency comparator and injects a small amount of current into the loop filter/integrator that causes the phase/frequency comparator to maintain an exactly balanced output. The result is that the reference spurs generated by the phase/frequency detector are vastly reduced. Figure 1 shows the block diagram for this approach. Circuit operation is as follows:

1. The output from the phase and frequency detector is differentially connected to the loop filter as in a conventional loop.
2. Each output from the phase and frequency detector is also connected to a pulse amplifier.
3. The output from each pulse amplifier is rectified and filtered.
4. The outputs from the detectors are differentially connected to an integrator. Any difference in the output voltage of the two detectors causes the integrator voltage to ramp to a new voltage forcing the difference to be eliminated.
5. The output from the integrator is summed into the loop filter amplifier. This results in virtually all of the offsets being cancelled, and the reference spurs being greatly reduced.

Figure 2 shows an actual implementation of the circuit as it was used in a frequency synthesizer. A detailed description on the operation is as follows:

1. Each output from the phase and frequency detector (an MC12040 in this case) is passed through an RC lowpass filter that acts as a pulse stretcher.
2. The output of each low pass filter goes two places
   A. To the loop filter
   B. To a pulse amplifier
3. Each pulse amplifier (Q1 and Q2 and their associated circuitry) is a discrete amplifier having a voltage gain of around 200. The actual gain is not important as long as it is reasonably large. This circuit amplifies any pulses present at the output of the phase detector to a level that can be detected by the diode detectors.
4. AC coupled to each pulse amplifier

output is a Schottky diode detector (D1, D2, and D3, D4) connected as a voltage doubler. Because in this case, the op-amp is operating with a single supply, one side of each detector is biased to approximately 1/2 the supply voltage by the voltage divider R6, R7 and bypassed C5. It is imperative that the same reference be used for both detectors to prevent differential offsets from occurring.

5. The output from each detector is connected to one input of the integrator (U1,C7, and C8) through R8 and R14 respectively. If the outputs from the two detectors are equal, the output of the integrator is constant. If they are unequal, the integrator output ramps up or down until they are.

6. The integrator output is connected to one input of the loop filter through a large value resistor, R16. This resistor's value must be sufficiently large that even if the integrator (U1) goes to a rail, the loop can still lock. This entire spur reduction circuit only functions when the loop is locked. Too small of a coupling resistor at the integrator output can result in the loop being unable to lock. Because the offsets are generally relatively small, R16 should be quite large (roughly 200 times) in value compared with the other two input resistors to the loop filter.

7. A resistor (R15) is connected between the same loop filter input as R16, and ground. This resistor causes a slight offset at the input to the loop filter, and is used to force the nominal output voltage of U4 to be at the midpoint of its operating range.

8. The time constant of R8/C7 and R14/C8 should be equal and long compared to the reciprocal of the loop bandwidth.

A circuit has been described that greatly reduces reference spurs in a phase locked loop caused by DC offsets in the loop filter amplifier and the phase detector. The circuit is relatively simple, requires no adjustments, is stable over the entire MIL temperature range, and has virtually no adverse effect on the loop dynamics. This technique allows huge reductions in the reference spurs to be achieved with virtually no unwanted side effects.

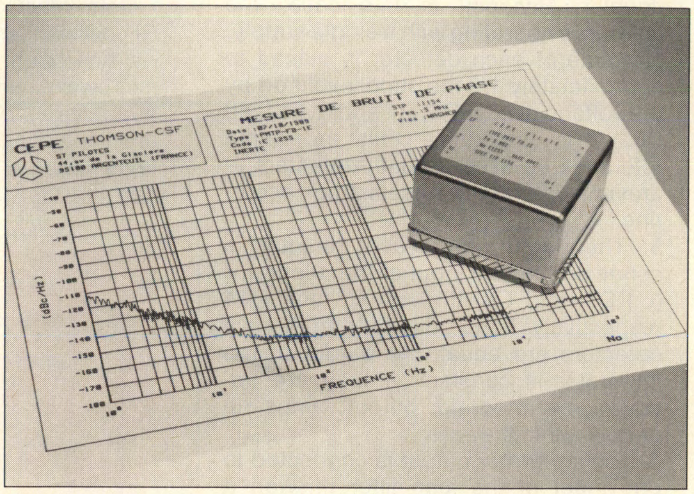
The circuit has been used in several UHF and microwave frequency synthesizers and has resulted in a sideband reduction of approximately 40 dB in all cases. The reduction in filter complexity and the improvement in loop dynamics was dramatic. *RF*

About the Authors

John MacConnell is manager of the RF division at Itron in Saratoga, CA. He has been working in the RF and microwave areas since 1971, with particular emphasis on frequency synthesis techniques and communications systems.

Richard Booth has been involved with the RF and microwave communications and navigation industry since 1973. He may be reached at Itron, 20520 Prospect Road, #2, Saratoga, CA 95070. His telephone number is (408) 973-7016.

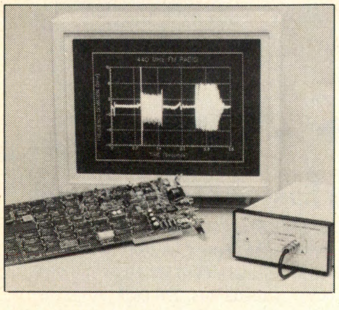
Ultra Stable Ovenized Crystal Oscillator



Thomson-CSF subsidiary, CEPE, introduces an ultra stable, low phase noise crystal ovenized oscillator, model PMTP FB, with very short warm up time, using a high performance SC crystal. The oscillator is miniaturized by using surface mount technology in a hermetic metal package. Standard frequencies are 5 MHz and 10 MHz, typically used in synthesizers/test equipment, ground based radar and ground based satellite stations. Frequency stability is better than $\pm$ 5 $\times$ 10^{-11} from −40 to +70 degrees C, and a typical noise floor value of −168 dBc/Hz. Minimum harmonic attenuation is −40 dBc. The output signal is a sinewave with a minimum of 400 mV rms (into 50 ohms). Frequency adjustment is by external potentiometer and/or control voltage. The oscillator meets MIL-STD 202 for non-operating shocks (Method 213 Test Condition A) and vibrations (Method 204 Test Condition A). The package size is 2.64" $\times$ 2.37" $\times$ 1.58" and the weight is 6.35 oz. Pricing at a 10 pc level is from \$970 to \$2430 depending upon stability specification, and availability is two months ARO.
Thomson-CSF Electron Tubes and Devices Corp. Special Products Division
INFO/CARD #200

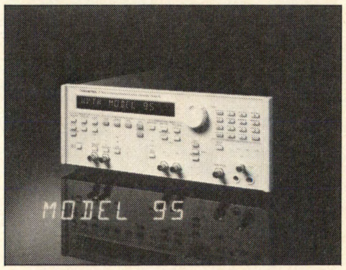
Digital Modulation Analyzer

The GT2210S Modulation Analysis System from Guide Technology Incorporated allows RF engineers to measure both low deviation FM and wide deviation frequency transients over a range of DC to 1.3 GHz. Measurement resolution can detect frequency deviations as small as 1 part per million with a 1 kHz sampling rate, and better resolution is possible at slower sampling rates. Typical applications include: testing UHF frequency hopping radio transmitters, testing VHF FM transceivers for peak deviation, testing RC, LC and crystal oscillators for long term drift and phase noise, or evaluating the transient response of VCOs and PLLs. Features of the GT2210S include: DC to 1.3 GHz frequency range, 10 digits per second frequency resolution, 10 mV sensitivity when measuring low level signals, and operates with XT, AT or 386 compatibles.
Guide Technology, Inc.
INFO/CARD #199



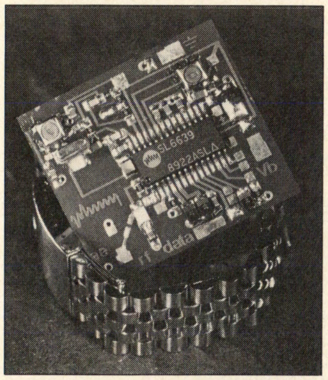
20 MHz Arbitrary Function and Trigger Generator

Model 95 synthesized 20 MHz arbitrary function generator with a built-in trigger generator has been announced by Wavetek San Diego, Incorporated. The generator features programmable sine, triangle, square, DC, and arbitrary waveforms. Variable symmetry allows the creation of pulse and ramp waveforms, and variable phase angle allows the output signal to be phase-shifted in relation to an external locking source. The frequency range is 1 to 20 MHz with 0.001 percent accuracy and user-selectable output impedances of 50, 75, 135, and 600 ohms. In addition, both digital and analog filters are provided for sine outputs. Model 95's advanced arbitrary has full 20 MHz sampling frequency and allows sentry, editing and storage of non-standard, user-defined waveshapes. Pricing starts at \$4,495 and delivery is four weeks ARO.
Wavetek San Diego, Inc.
INFO/CARD #197



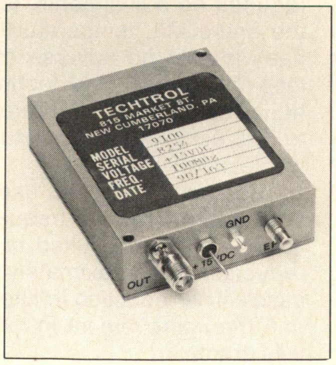
RF Data Receiver

Model SL6639 RF data receiver from Plessey Semiconductor is designed for applications in frequencies up to 200 MHz and data rates of more than 1.2 Kb. It has on-chip channel filters reducing the need for additional components. This cascadable amplifier and data demodulator is designed to detect FSK modulated data. Gyrator filters have been utilized to provide adjacent channel rejection in excess of 70 dB at 25 kHz channel spacing. Alignment of the channel filters requires only one potentiometer. Sensitivity is typically −124 dBm at 150 MHz from a 50 ohm source, and stability is enhanced by using a cascadable RF configuration. Model SL6639 is ideal for use in modems, security devices and pagers. Pricing is \$7.27 in quantities of 100,000.
Plessey Semiconductors Inc.
INFO/CARD #196



Ultra Low Noise Crystal Oscillator

Techtrol Cyclonetics has released an ultra low noise crystal oscillator. The 9000 series is available between 40 and 125 MHz, delivering an output level of +20 dBm and holding a frequency stability of $\pm$ 0.001 percent between −20 to +60 degrees Celsius. The oscillator offers SSB phase noise of −98 dBc/Hz at 10 Hz, −128 dBc/Hz at 100 Hz, −156 dBc/Hz at 1 kHz, −170 dBc/Hz at 10 kHz, and −173 dBc/Hz at 100 kHz. Signal harmonics are −26 dBc and spurious are −90 dBc. AFC capability is $\pm$ 15 PPM with $\pm$ 6 volts control. The RF power output for the 9000 series is +20 dBm minimum and operating voltage is +15 VDC. The unit is housed in an aluminum 2.5" $\times$ 2.0" $\times$ 0.68" package. Connectors for the oscillator are type SMA-F.
Techtrol Cyclonetics, Inc.
INFO/CARD #195



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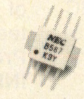
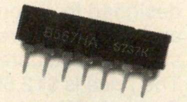
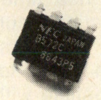
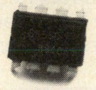
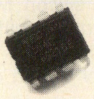
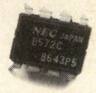
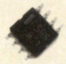
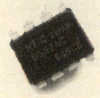
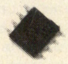
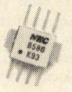
Frequency Dividers — **NEC**

÷ 2  UPG502 UPB565 UPG504 UPB581 UPB558 UPB584	**÷ 4**  UPG501 UPB565 UPG503 UPB582 UPB585	**÷ 8**  UPG506 UPB565 UPB567	**÷ 2/4/8 ÷ 8/9**  UPB555 UPB587	**÷ 10/11**  UPB551 UPB552 UPB554
÷ 16/17  UPB553 UPB556 UPB555 UPB571	**÷ 20/21**  UPB572	**÷ 20/22**  UPB551 UPB552 UPB554	**÷ 32/33**  UPB555 UPG569 UPB571	**÷ 40/41**  UPB572
÷ 40/44  UPB551 UPB552 UPB554	**÷ 64**  UPB564 UPB565	**÷ 64/65**  UPB566 UPB569 UPB571	**÷ 64/68**  UPB562 UPB568	**÷ 80/81**  UPB572
÷ 128 ÷ 128/129  UPB564 UPB566	**÷ 128/64**  UPB588	**÷ 128/136**  UPB562 UPB568	**÷ 256**  UPB564	**÷ 256/512**  UPB586

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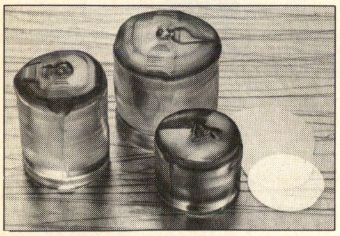
Lithium Niobate

Crystal Technology, Inc. has made available lithium niobate single crystals for all acoustical and optical applications for the



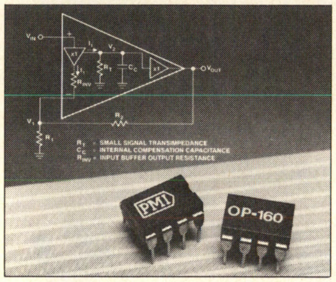
material. The crystals enable the establishment of new industry standards for lithium niobate uniformity, relevant to all acoustical and optical applications for the material.
Crystal Technology, Inc.
INFO/CARD #194

Microstrip Test Station

The MTF26 provides two-port measurement capacity for MIC structures, from DC to 26.5 GHz. The coax-to-microstrip launchers can accommodate test structures with offset ports and substrates as small as 120 mils or as large as 4 inches square. The fixture is designed to accommodate MMIC and MIC structures on bare substrate or mounted on carriers. The price is $12,500.
Cascade Microtech
INFO/CARD #193

Current Feedback Op Amp

Precision Monolithics Incorporated has released the OP-160, a very high speed current feedback operational amplifier that is capa-



ble of driving large capacitive loads without oscillation. The OP-160 has gain bandwidths exceeding 400 MHz. In unity-gain applications, it has a typical slew rate of over 1300 V/μs and is guaranteed to exceed 1000 V/μs.
Precision Monolithics
INFO/CARD #192

Oscillator Tester

Model 1010 from PRA, Inc. measures the waveform parameters of TTL and CMOS clock oscillators. Measurements include frequency, rise time, fall time, duty cycle, supply current, and logic levels. It covers frequencies between 1 kHz to 100 MHz and is compatible with PC XT or AT computers.
PRA, Inc.
INFO/CARD #191

VGC Amplifiers

Avantek, Incorporated has introduced two new silicon MMIC variable-gain-control amplifiers, IVA-05218 (1.5 GHz) and IVA-04118 (2.5 GHz) for use in communications systems. The IVA-05218 provides up to 30 dB of power gain controllable over a 30 dB range, and the IVA-04118 provides 27 dB of power gain controllable over a 35 dB range.
Avantek, Inc.
INFO/CARD #190

GPS Time and Frequency Receiver

The AUSTRON Model 2201 GPS time and frequency receiver tracks as many as eight satellites at one time. It tracks the C/A code on the L1 carrier frequency (1575.42 MHz) carrier and generates a 1 PPS and 1 PPM signal synchronized to within 100 ns of Universal Time Coordinated (UTC) and provides time and frequency calibration with the US Naval Observatory.
Austron, Inc.
INFO/CARD #189

100 to 2000 MHz Surface-Mount Amplifier

The WJ-SMA36 surface-mount

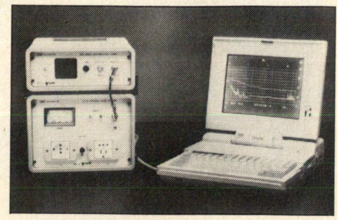
amplifier from Watkins-Johnson Company operates over the 100 to 2000 MHz frequency range and the 0 to 50 degree centigrade temperature range. Specifications include: 15.5 dB minimum gain, 7.0 dB maximum noise figure and 11.0 dBm minimum power output at 1 dB compression.
Watkins-Johnson
INFO/CARD #188

Hybrid VCXO

A voltage controlled crystal oscillator, the VH2340, is now available from Murata Erie North America. This VCXO is available in the frequency range of 15 MHz to 50 MHz and features wide frequency versus control voltage deviation capability of ± 100 ppm, as V_c is varied from 0 to − 5.0 VDC.
Murata Erie North America
INFO/CARD #187

Conducted Interference Test System

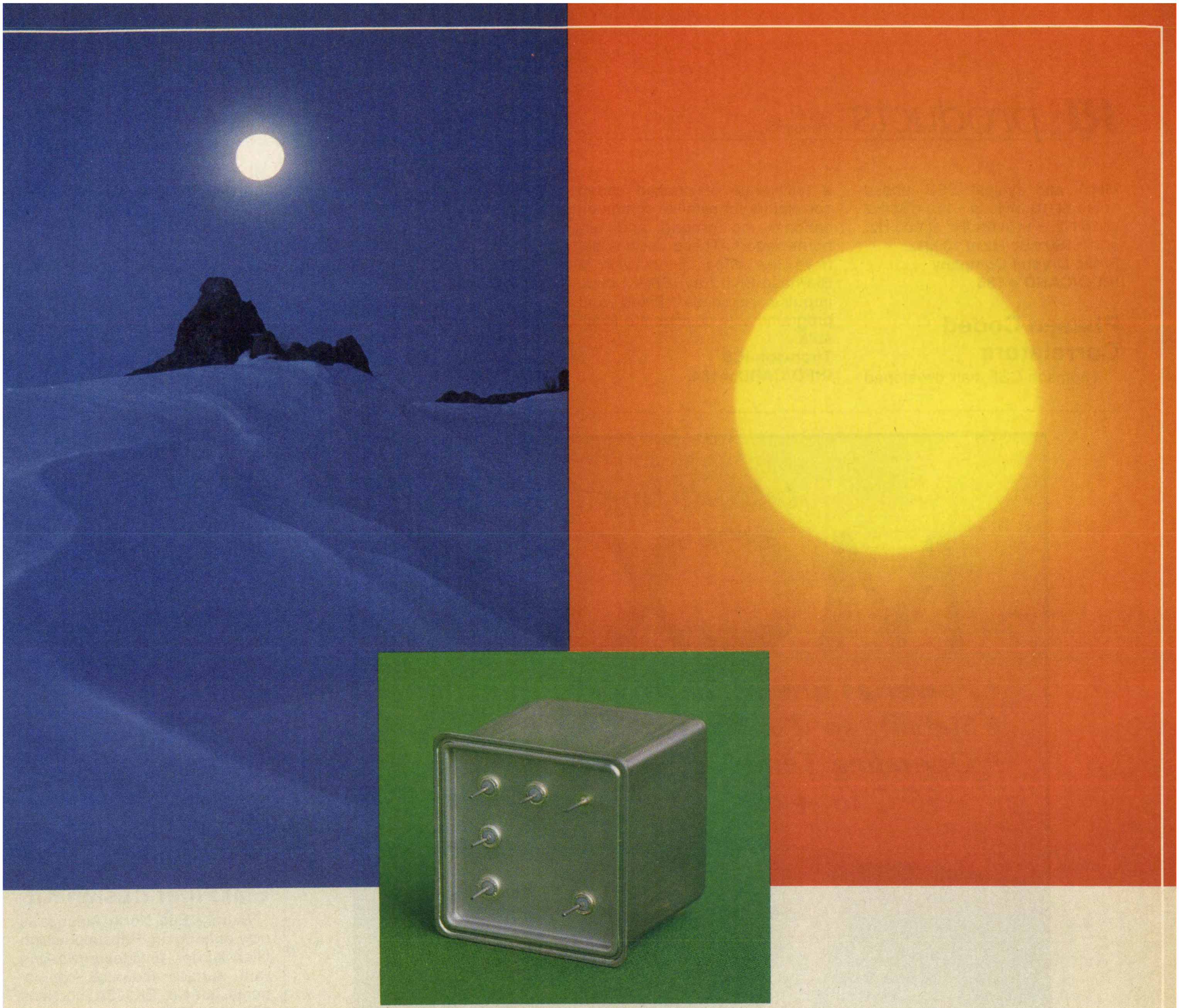
The Vianello PMM-8010 Meas-



uring System from IBEX measures EMI for FCC, DOC, VDE, and other CISPR conforming standards. Comprised of a receiver, three separate Line Impedance Networks (LISNs), and a half slot PC interface card, measurements can be taken in either manual or automatic modes.
IBEX Group Inc.
INFO/CARD #186

Low Profile Ovenized Oscillator

Piezo Crystal Company announces the availability of Model 2900082. This oscillator was designed for commercial/military communication, instrumentation, and GPS applications. The frequency range is from 5 to 15



Shown actual size. 1.3 in. x 1.3 in. x 1.3 in.

Neither snow nor sleet nor freezing winds nor scorching heat can keep the EMXO from its appointed rounds.

There is nothing on the market today that can match Ball, Efratom's EMXO world-class series sub-miniature military quartz crystal oscillator in size, versatility, stability and reliability.

A technological breakthrough by Efratom engineers in resonator design concepts, coupled with major advances in package design, has resulted in the smallest, most reliable quartz oscillator available today.

Designed for low power, low volume applications, the EMXO can be direct-mounted on a PC board or card. The **10 MHz and the 10.23 MHz**, sinusoidal or TTL output, provide outstanding short- and long-term stability over an extreme range of environmental conditions, handling temperature extremes of **−55 C to +95 C**.

The EMXO is cost-effective for critical time and frequency programs. It is especially advantageous when utilized in mobile/transportable and portable applications where **fast warm-up, low power consumption and small volume** are required.

Ideal applications for the EMXO include digital communications, radar, precision measurement instruments, time-code generators, geophysical survey positioning and navigation systems, satellite tracking and guidance control, and standard time and time interval generation and transfer.

Given its extremely small size, **1.3 inches cubed**, the EMXO offers an outstanding **MTBF >100,000 hours**.

For the past 18 years, Ball, Efratom has been the industry leader in rubidium atomic oscillators. Now it plans to lead the industry in **high-precision SC cut quartz crystal oscillators**.

For more information on Efratom's problem-solving EMXO quartz oscillator series, write or call:

Ball
Efratom Division

3 Parker
Irvine, California 92718-1605
Telephone (714) 770-5000
Telex 685-635
Fax (714) 770-2463

MHz, and typical SSB phase noise at 10 MHz is −110 dBc/Hz at 10 Hz, −140 dBc/Hz at 100 Hz, and −160 dBc/Hz at 10 kHz.
Piezo Crystal Company
INFO/CARD #185

Phased Coded Correlators

Thomson-CSF has developed a full range of phased coded correlators for satellite communications and ground and airborne radars. These devices behave like linear filters with a Bi-Phase Shift Key (BPSK) finite impulse response. Fixed and programmable devices are available.
Thomson-ICS
INFO/CARD #184

ECL Clock to 700 MHz

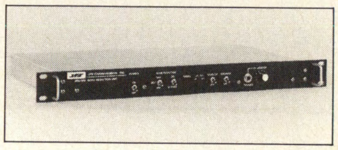
Clock oscillator Model CO-233KEQ provides complementary sub-nanosecond ECL logic compatible outputs at any specified frequency in the 150-700 MHz frequency range. The output is derived from either a 100K ECL or ECLiPS gate, depending on the output frequency.
Vectron Laboratories
INFO/CARD #183

14-Bit Analog/Digital Converter

Analog Devices' AD9014, gives encode rates up to 10 megasamples per second. The AD9014 has input frequencies of 504 kHz, .3 MHz. At either frequency, minimum spurious-free dynamic range (SFDR) is 90 dB, decreasing to 86 dB with input frequencies up to 4.3 MHz.
Analog Devices
INFO/CARD #182

Noise Reduction Unit

The NRU-500 uses DSP algorithms to adjust itself to the



characteristics of the input noise, allowing it to suppress broadband noise, repetitive impulse noise, time-varying noises, and interfering heterodynes.
JPS Communications, Inc.
INFO/CARD #181

EMI Filter Design Kit

Murata Erie North America is now offering an EMI suppression filter kit for engineers working with surface mounted components. Kit No. EK115A, contains 220 chip EMI filter components including solid ferrite devices and the NFM series of EMI filters.
Murata Erie North America
INFO/CARD #180

500 MHz Amplifier-Attenuator Module

A new 4 channel amplifier/attenuator VME module designed for use with the Series 2000 Multichannel Waveform Sampling System has been announced by Analytek/Tektronix. The Model 2004A module features DC to 500 MHz bandwidth with 100 μV sensitivity at 50 megaohms input impedance.
Analytek/Tektronix
INFO/CARD #178

High Temperature Schottky Mixer Diode Chips

These new diode chips from

FEI Microwave feature a 300 degree Celsius burn-in capability, and can withstand a 200 degree Celsius long term storage temperature. Specifications include less than 6 dB noise figure, breakdown voltage greater than 5 volts, capacitance less than 0.08 pF and current leakage less than 1.0 nA at 3 volts.
FEI Microwave, Inc.
INFO/CARD #179

Silicon Hyperabrupt Varactors

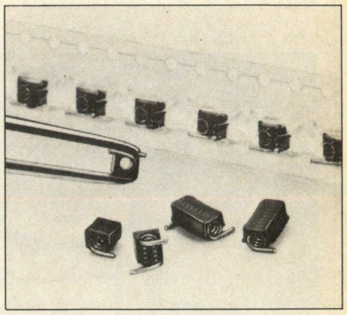
Alpha Industries has released three silicon hyperabrupt varactor diodes. Series SMV1204, SMV2204, and DKV6504 are come in a variety of SMT packages and allow a capacitance change greater than 7:1 over a voltage range of less than 9 volts, without sacrificing the diode Q.
Alpha Industries
INFO/CARD #177

GaAs SPST

The model DSO841 GaAs SPST switch has a transition time of 680 ps and uses less than 2 mA of current from a single +5 VDC supply. This device operates from 10 to 200 MHz with 1.5 dB insertion loss and 72 dB isolation up to 100 MHz, 60 dB isolation up to 200 MHz. The total switching speed is less than 7 ns.
Daico Industries, Inc.
INFO/CARD #175

SMT Air Core Inductors

Coilcraft is introducing a line of inductors that range from 2.5 to 43 nH with minimum Q values of 100 to 145 at 150 MHz. Tolerance is ± 5 percent and self resonance is typically greater than 3 GHz.
Coilcraft
INFO/CARD #176



RF software

Electromagnetics Software and Hardware

The Data Transport Computer™ and EM WAVETRACER™ have been developed by Wavetracer Inc. to solve complex problems in physical phenomena including electromagnetic scattering in antenna design, design of low-observable vehicles, radar design, and sensor design. EM WAVETRACER is based on a Finite Difference Time Domain (FDTD) method developed by MIT, and the DTC is a parallel computer designed for data processing, transporting, and visualization on problems of two or more dimensions. The DTC attaches to most industry-standard workstations.
Wavetracer Inc.
INFO/CARD #215

MMIC Design Program

MMIC_CAD, Inc. offers a new MMIC design program named MMIC_key. MMIC_key will calculate the inductance, capacitance, high frequency resistance and the microstrip characteristics of any number of signal paths in a MMIC layout to ±10 percent. This program can be configured to write a file containing the nodal netlist representation of the results that can be directly included in a SPICE or Touchstone™ netlist file. MMIC_key will run on any IBM AT or compatible with DOS 2.0 or higher. An EGA or VGA monitor, a math coprocessor chip, and at least 512k of RAM are required. The price is \$995 plus shipping and handling.
MMIC_CAD, Inc.
INFO/CARD #214

RF Circuit Software

RFSynthesist™ from ingSOFT, Ltd. works as a stand-alone application, as well as a part of the RFDesigner™ system. RFSynthesist includes: filter synthesis, (Butterworth, Chebyshev, elliptic, coupled resonators); calculation of transmission line characteristics; synthesis of microstrip lines; and calculation of coupled transmission line characteristics in stripline and microstrip configurations. RFSynthesist runs on all Macintosh models.
ingSOFT, Ltd.
INFO/CARD #213

Digital Signal Processing System Design

Comdisco Systems introduces the Hardware Design System™ option for the Signal Processing WorkSystem™ which lets engineers account for the numerical precision of real world hardware when designing digital signal processing systems. Design formats include VHDL, LSIM, or Verilog EDIF. Priced at \$10,000, HDS is available on Sun-3™ and Sun-4™ (SPARC), Hewlett-Packard/Apollo® DN3000 and 4000 series, and DECstation™ 2100 and 3100 series.
Comdisco Systems, Inc.
INFO/CARD #212

Split-Tee Power Divider

By Stanislaw Rosloniec and Piotr Lochowicz
Dept. of Electronic Engineering
Warsaw Technical University

Some methods of evaluating split-tee power dividers have drawbacks which limit their usefulness in the analysis. This paper introduces an alternative approach for analyzing split-tee power dividers and uses a computer program to simplify the math.

Figure 1 presents the electrical scheme of the improved split-tee power divider described by Parad and Moynihan in their classical paper (1). Dividers (combiners) of this type are used for various applications, especially at RF and microwave frequencies. Hence, it is necessary to analyze their frequency properties.

In the general case, the divider shown in Figure 1 is asymmetric with respect to the horizontal plane x - x', and for this reason its analysis cannot be done by using the even- and odd- mode excitation method (2,3,4). In addition, the analysis technique suggested in (1) is not suitable for this purpose since the auxiliary terminating impedances Z_{a2} and Z_{a3} (see Figure 1) are complex and dependent upon frequency. Therefore, in the present contribution a new numerical algorithm is presented which allows us to analyze the divider under consideration without difficulty. An example of this algorithm is presented below.

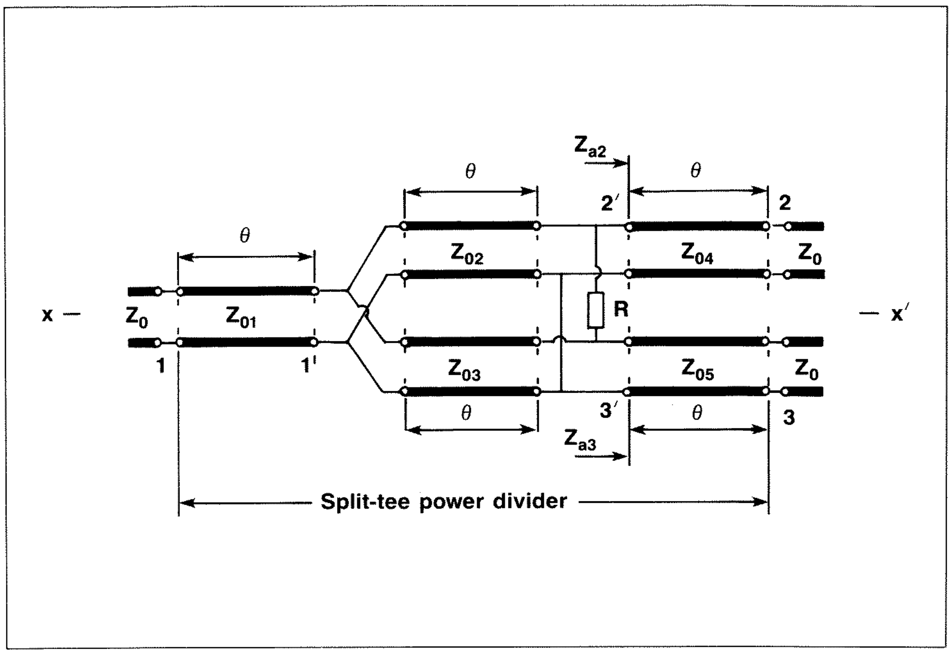


Figure 1. Electrical scheme of the improved unequal split-tee power divider.

Circuit Transformations

At any electrical length, θ, the main divider section (see Figure 2) may be treated as a linear three-port network characterized by the following admit-

$$\begin{bmatrix} I_1' \\ I_2' \\ I_3' \end{bmatrix} = \begin{bmatrix} Y_{11}, Y_{12}, Y_{13} \\ Y_{21}, Y_{22}, Y_{23} \\ Y_{31}, Y_{32}, Y_{33} \end{bmatrix} \begin{bmatrix} U_1' \\ U_2' \\ U_3' \end{bmatrix} \quad (1)$$

Where

$$Y_{11} = -j\left(\frac{Y_{02}}{\tan\theta} + \frac{Y_{03}}{\tan\theta}\right) \quad (2)$$

$$Y_{12} = Y_{21} = j\,\frac{Y_{02}}{\sin\theta} \quad (3)$$

$$Y_{13} = Y_{31} = j\,\frac{Y_{03}}{\sin\theta} \quad (4)$$

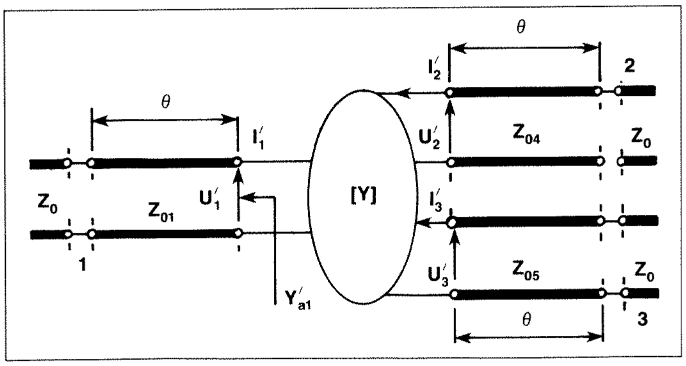


Figure 2. Definition of admittance parameters[Y].

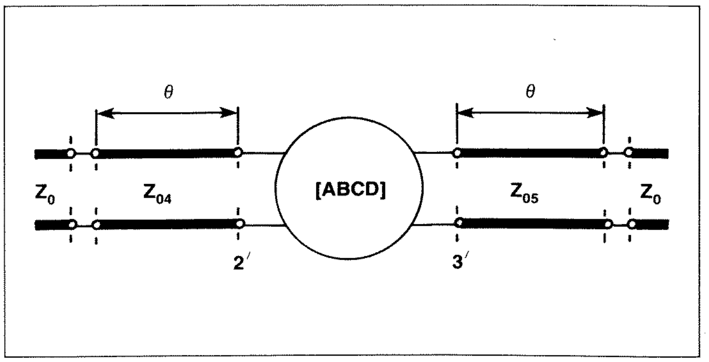


Figure 3. Equivalent circuit for determining S_{22}, S_{23}, S_{32}, and S_{33}.

Scattering parameters	$k^2 = 2$		
	0 = 1.0 rad	0 = 1.5 rad	0 = 2.0 rad
S_{11}	0.1431 < 1.2157	0.0049 < −0.8164	0.0906 < −0.8121
S_{12}	0.5571 < −2.9925	0.5770 < −4.4964	0.5653 < 0.2695
S_{13}	0.8136 < −2.9567	0.8166 < −4.4963	0.8171 < 0.2538
S_{21}	0.5571 < −2.9925	0.5770 < −4.4964	0.5653 < 0.2695
S_{22}	0.2104 < −0.3836	0.0317 < −1.4168	0.1705 < 0.6647
S_{23}	0.1629 < −3.5006	0.0139 < −4.6038	0.1092 < −2.4048
S_{31}	0.8136 < −2.9567	0.8166 < −4.4963	0.8171 < 0.2538
S_{32}	0.1629 < −3.5006	0.0139 < −4.6038	0.1092 < −2.4048
S_{33}	0.1203 < 1.7402	0.0036 < −5.3061	0.0804 < 4.8973

Table 1. Scattering parameters.

$$Y_{22} = \frac{1}{R} - j\,\frac{Y_{02}}{\tan\theta} \quad (5)$$

$$Y_{23} = Y_{32} = -\frac{1}{R} \quad (6)$$

$$Y_{33} = \frac{1}{R} - j\,\frac{Y_{03}}{\tan\theta} \quad (7)$$

$$Y_{02} = \frac{1}{Z_{02}} \quad (8)$$

$$Y_{03} = \frac{1}{Z_{03}} \quad (9)$$

Due to description (1) the scattering parameters of the entire divider can be easily obtained by using the conventional approach, i.e. the transfer matrix technique (5,6). As an example, let us calculate the elements S_{22}, S_{23}, S_{32} and S_{33} of the scattering matrix being sought. In this case the equivalent two-port network placed between planes 2' and 3' is similar to that shown in Figure 3. According to (1) and Figure 2 the transfer matrix (ABCD) of this network is

$$A = -\frac{Ye_{22}}{Ye_{21}} \quad (10)$$

$$B = -\frac{1}{Ye_{21}} \quad (11)$$

$$C = Ye_{12} - \frac{Ye_{11}Ye_{22}}{Ye_{21}} \quad (12)$$

$$D = -\frac{Ye_{11}}{Ye_{21}} \quad (13)$$

where

$$Ye_{11} = Y_{22} - \frac{Y_{12}Y_{21}}{Y_{11} + Y'_{a1}} \quad (14)$$

$$Ye_{12} = Y_{23} - \frac{Y_{13}Y_{21}}{Y_{11} + Y'_{a1}} \quad (15)$$

$$Ye_{21} = Y_{32} - \frac{Y_{12}Y_{31}}{Y_{11} + Y'_{a1}} \quad (16)$$

$$Ye_{22} = Y_{33} - \frac{Y_{13}Y_{31}}{Y_{11} + Y'_{a1}} \quad (17)$$

$$Y'_{a1} = Y_{01}\left(\frac{Y_0 + jY_{01}\tan\theta}{Y_{01} + jY_0\tan\theta}\right) \quad (18)$$

$$Y_0 = \frac{1}{Z_0} \quad (19)$$

$$Y_{01} = \frac{1}{Z_{01}} \quad (20)$$

Consequently, the transfer matrix of the entire divider may be calculated by multiplying the corresponding transfer matrices of the cascade components shown in Figure 3. When the resulting matrix is known, then we can evaluate the divider related to its scattering matrix, i.e. the scattering elements mentioned above. It is obvious that the remaining scattering parameters of the divider being analyzed may be evaluated in a similar manner. Such an approach has been used in the computer program UPD which is an integral part of this paper. Some illustrative results obtained by means of this program are summarized in Table 1.

The program UPD also calculates the voltage standing wave ratios $VSWR_1$, $VSWR_2$, $VSWR_3$, couplings C_{12}, C_{13}, and isolation I_{23}, which are related to the scattering parameters as follows:


```
10 REM  UPD
20 CLS: DEFDBL A-Z: DEFINT I,J
30 DIM M(3,3), V(3,3)
40 PRINT "UNEQUAL POWER DIVIDER": PRINT
50 PRINT
60 PRINT "This program is an integral part"
70 PRINT "of the paper * Split-tee power "
80 PRINT "divider (supplement) * by Stanislaw"
90 PRINT "Rosloniec and Piotr Lochowicz."
100 PRINT
110 PRINT "Data :"
120 PRINT
130 INPUT "Zo[ohm] = ",ZO
140 INPUT "k^2 = ",KK
150 INPUT "O[rad]  = ",O
160 PRINT
170 IF ZO<=0 THEN PRINT "Error : Zo <= 0": GOTO 120
180 IF KK<=0 THEN PRINT "Error : k^2 <= 0": GOTO 120
190 IF O<=0 THEN  PRINT "Error : O <= 0": GOTO 120
200 PRINT "Results, see Fig. 1"
210 PRINT
220 LET PI=3.141592653#
230 LET K=SQR(KK)
240 FOR I=1 TO 3
250 FOR J=1 TO 3
260 LET M(I,J)=0: LET V(I,J)=0
270 NEXT J
280 NEXT I
290 LET ZO1=ZO*SQR(SQR(K/(1+KK)))
300 LET ZO2=ZO*SQR(SQR(KK*K*(1+KK)))
310 LET ZO3=ZO*SQR(SQR((1+KK)/(KK*KK*K)))
320 LET ZO4=ZO*SQR(K)
330 LET ZO5=ZO/SQR(K)
340 LET R=ZO*(1+KK)/K
350 PRINT
360 PRINT "Electrical parameters, see Fig.1"
370 PRINT
380 PRINT "Zo1[ohm] ="USING "####.####";ZO1
390 PRINT "Zo2[ohm] ="USING "####.####";ZO2
400 PRINT "Zo3[ohm] ="USING "####.####";ZO3
410 PRINT "Zo4[ohm] ="USING "####.####";ZO4
420 PRINT "Zo5[ohm] ="USING "####.####";ZO5
430 PRINT
440 PRINT "R[ohm] ="USING "####.####";R
450 PRINT
460 IF ABS(O-PI/2)<.000001 THEN GOTO 1930
470 IF ABS(SIN(O))<.000001 THEN GOTO 2000
480 LET C=COS(O)
490 LET S=SIN(O)
500 LET T=TAN(O)
510 REM Admittance parameters
520 LET B11=-1/(ZO2*T)-1/(ZO3*T): LET G11=0
530 LET B12=1/(ZO2*S): LET G12=0
540 LET B13=1/(ZO3*S): LET G13=0
550 LET B22=-1/(ZO2*T): LET G22=1/R
560 LET B23=0: LET G23=-1/R
570 LET B33=-1/(ZO3*T)
580 REM Admittances Ya1, Ya2 and Ya3
590 LET YX=1/ZO1
600 GOSUB 2110

610 LET GA1=GI:LET BA1=BI
620 LET YX=1/ZO4
630 GOSUB 2110
640 LET GA2=GI: LET BA2=BI
650 LET YX=1/ZO5
660 GOSUB 2110
670 LET GA3=GI: LET BA3=BI
680 REM  (1-2) ABCD matrix
690 LET GII=G22: LET BII=B33
700 LET GAI=GA3: LET BAI=BA3
710 LET GKL=G11: LET BKL=B11
720 LET GMN=G13: LET BMN=B13
730 LET GPQ=G13: LET BPQ=B13
740 GOSUB 2160
750 LET GE11=GEKL: LET BE11=BEKL
760 LET GKL=G12: LET BKL=B12
770 LET GPQ=G23: LET BPQ=B23
780 GOSUB 2160
790 LET GE12=GEKL: LET BE12=BEKL
800 LET GE21=GEKL: LET BE21=BEKL
810 LET GKL=G22  : LET BKL=B22
820 LET GMN=G23  : LET BMN=B23
830 GOSUB 2160
840 LET GE22=GEKL: LET BE22=BEKL
850 GOSUB 2250
860 LET AR2=C: LET AI2=0
870 LET BR2=0: LET BI2=ZO4*S
880 LET CR2=0: LET CI2=S/ZO4
890 LET DR2=C: LET DI2=0
900 GOSUB 2380
910 LET AR1=C: LET AI1=0
920 LET BR1= 0: LET BI1=ZO1*S
930 LET CR1=0: LET CI1=S/ZO1
940 LET DR1=C: LET DI1=0
950 GOSUB 2380
```

Figure 4. Computer Program UPD.

$$VSWR_1 = \frac{1 + |S_{11}|}{1 - |S_{11}|} \tag{21}$$

$$VSWR_2 = \frac{1 + |S_{22}|}{1 - |S_{22}|} \tag{22}$$

$$VSWR_3 = \frac{1 + |S_{33}|}{1 - |S_{33}|} \tag{23}$$

$$C_{12} = 20 \log \left(\frac{1}{|S_{12}|}\right), \text{ dB} \tag{24}$$

$$C_{13} = 20 \log \left(\frac{1}{|S_{13}|}\right), \text{ dB} \tag{25}$$

$$I_{23} = 20 \log \left(\frac{1}{|S_{23}|}\right), \text{ dB} \tag{26}$$

```
960 REM  (1-2) Scattering parameters
970 GOSUB 2530
980 LET M(1,1)=M11: LET V(1,1)=V11
990 LET M(1,2)=M12: LET V(1,2)=V12
1000 LET M(2,1)=M21: LET V(2,1)=V21
1010 LET M(2,2)=M22: LET V(2,2)=V22
1020 REM  (1-3) ABCD matrix
1030 LET GII=G22: LET BII=B22
1040 LET GAI=GA2: LET BAI=BA2
1050 LET GKL=G11: LET BKL=B11
1060 LET GMN=G12: LET BMN=B12
1070 LET GPQ=G12: LET BPQ=B12
1080 GOSUB 2160
1090 LET GE11=GEKL: LET BE11=BEKL
1100 LET GKL= G13: LET BKL=B13
1110 LET GPQ=G23: LET BPQ=B23
1120 GOSUB 2160
1130 LET GE12=GEKL: LET BE12=BEKL
1140 LET GE21=GEKL: LET BE21=BEKL
1150 LET GKL=G22: LET BKL=B33
1160 LET GMN=G23: LET BMN=B23
1170 GOSUB 2160
1180 LET GE22=GEKL: LET BE22=BEKL
1190 GOSUB 2250
1200 LET AR2=C: LET AI2=0

1210 LET BR2=0: LET BI2=ZO5*S
1220 LET CR2=0: LET CI2=S/ZO5
1230 LET DR2=C: LET DI2=0
1240 GOSUB 2380
1250 LET AR1=C: LET AI1=0
1260 LET BR1=0: LET BI1=ZO1*S
1270 LET CR1=0: LET CI1=S/ZO1
1280 LET DR1=C: LET DI1=0
1290 GOSUB 2380
1300 REM  (1-3) Scattering parameters
1310 GOSUB 2530
1320 LET M(1,3)=M12: LET V(1,3)=V12
1330 LET M(3,1)=M21: LET V(3,1)=V21
1340 LET M(3,3)=M22: LET V(3,3)=V22
1350 REM  (2-3) ABCD matrix
1360 LET GII=G11: LET BII=B11
1370 LET GAI=GA1: LET BAI=BA1
1380 LET GKL=G22: LET BKL=B22
1390 LET GMN=G12: LET BMN=B12
1400 LET GPQ=G12: LET BPQ=B12
1410 GOSUB 2160
1420 LET GE11=GEKL: LET BE11=BEKL
1430 LET GKL=G23: LET BKL=B23
1440 LET GMN=G13: LET BMN=B13
1450 GOSUB 2160
1460 LET GE12=GEKL: LET BE12=BEKL
1470 LET GE21=GEKL: LET BE21=BEKL
1480 LET GKL=G22: LET BKL=B33
1490 LET GPQ=G13: LET BPQ=B13
1500 GOSUB 2160
1510 LET GE22=GEKL: LET BE22=BEKL
1520 GOSUB 2250
1530 LET AR2=C: LET AI2=0
1540 LET BR2=0: LET BI2=ZO5*S
1550 LET CR2=0: LET CI2=S/ZO5
1560 LET DR2=C: LET DI2=0
1570 GOSUB 2380
1580 LET AR1=C: LET AI1=0
1590 LET BR1=0: LET BI1=ZO4*S
1600 LET CR1=0: LET CI1=S/ZO4
1610 LET DR1=C: LET DI1=0
1620 GOSUB 2380
1630 REM  (2-3) Scattering parameters
1640 GOSUB 2530
1650 LET M(2,3)=M12: LET V(2,3)=V12
1660 LET M(3,2)=M21: LET V(3,2)=V21
1670 PRINT
1680 PRINT "Scattering parameters"
1690 PRINT
1700 FOR I=1 TO 3
1710 FOR J=1 TO 3
1720 PRINT "S (";I*10+J;") ="USING "###.####";M(I,J);:
  PRINT " exp( j";USING "###.####";V(I,J);: PRINT " )"
1730 NEXT J
1740 NEXT I
1750 PRINT
1760 INPUT "Enter 'CONT':",N$
1770 IF N$="CONT" THEN GOTO 1790
1780 GOTO 1750
1790 PRINT

1800 PRINT "Matching, coupling and isolation data"
1810 PRINT
1820 PRINT "VSWR(1) ="USING"##.####";(1+M(1,1))/(1-M(1,1))
1830 PRINT "VSWR(2) ="USING"##.####";(1+M(2,2))/(1-M(2,2))
1840 PRINT "VSWR(3) ="USING"##.####";(1+M(3,3))/(1-M(3,3))
1850 PRINT
1860 PRINT "C(1,2)[dB] ="USING"##.####";20/LOG(10)*LOG(1/M(1,2))
1870 PRINT "C(1,3)[dB] ="USING"##.####";20/LOG(10)*LOG(1/M(1,3))
1880 PRINT
1890 IF M(2,3)<.00001 THEN PRINT "I(2,3)[dB] > 100": GOTO 1910
1900 PRINT "I(2,3)[dB] = "USING"##.####";20/LOG(10)*LOG(1/M(2,3))
1910 PRINT
1920 END
1930 REM Procedure 1930
1940 LET M(1,2)=SQR(1/(1+KK)): LET V(1,2)=PI/2
1950 LET M(2,1)=M(1,2): LET V(2,1)=PI/2
1960 LET M(1,3)=K/SQR(KK+1)
1970 LET M(3,1)=M(1,3)
1980 LET V(1,3)=PI/2: LET V(3,1)=PI/2
1990 GOTO 1700
2000 REM Procedure 2000
2010 LET M(1,1)=1/3: LET V(1,1)=PI
2020 LET M(1,2)=2/3: LET V(1,2)=-3*O
2030 LET M(1,3)=2/3: LET V(1,3)=-3*O
2040 LET M(2,1)=2/3: LET V(2,1)=-3*O
2050 LET M(2,2)=1/3: LET V(2,2)=PI
2060 LET M(2,3)=2/3: LET V(2,3)=-4*O
2070 LET M(3,1)=2/3: LET V(3,1)=-3*O
2080 LET M(3,2)=2/3: LET V(3,2)=-4*O
2090 LET M(3,3)=1/3: LET V(3,3)=PI
2100 GOTO 1700
2110 REM Subroutine 2110
2120 LET D=YX*YX+T*T/(ZO*ZO)
2130 LET GI=YX*YX*(1+T*T)/(ZO*D)
2140 LET BI=YX*T*(YX*YX-1/(ZO*ZO))/D
2150 RETURN
2160 REM Subroutine 2160
2170 LET GXN=GMN*GPQ-BMN*BPQ
2180 LET BXN=BMN*GPQ+GMN*BPQ
2190 LET GXD=GII+GAI
2200 LET BXD=BII+BAI
2210 LET D=GXD*GXD+BXD*BXD
2220 LET GEKL=GKL-(GXN*GXD+BXN*BXD)/D
2230 LET BEKL=BKL-(GXD*BXN-GXN*BXD)/D
2240 RETURN
2250 REM Subroutine 2250
2260 LET D=GE21*GE21+BE21*BE21
2270 LET AR1=(-GE21*GE22-BE21*BE22)/D
2280 LET AI1=(BE21*GE22-GE21*BE22)/D
2290 LET BR1=-GE21/D
2300 LET BI1=BE21/D
2310 LET GXN=GE11*GE22-BE11*BE22
2320 LET BXN=BE11*GE22+GE11*BE22
2330 LET CR1=GE12-(GXN*GE21+BXN*BE21)/D
2340 LET CI1=BE12-(BXN*GE21-GXN*BE21)/D
2350 LET DR1=(-GE21*GE11-BE21*BE11)/D
2360 LET DI1=(BE21*GE11-GE21*BE11)/D
2370 RETURN
2380 REM Subroutine 2380
2390 LET AR=AR1*AR2-AI1*AI2+BR1*CR2-BI1*CI2

2400 LET AI=AR1*AI2+AI1*AR2+BR1*CI2+BI1*CR2
2410 LET BR=AR1*BR2-AI1*BI2+BR1*DR2-BI1*DI2
2420 LET BI=AI1*BR2+AR1*BI2+BI1*DR2+BR1*DI2
2430 LET CR=CR1*AR2-CI1*AI2+DR1*CR2-DI1*CI2
2440 LET CI=CR1*AI2+CI1*AR2+DI1*CR2+DR1*CI2
2450 LET DR=CR1*BR2-CI1*BI2+DR1*DR2-DI1*DI2
2460 LET DI=CR1*BI2+CI1*BR2+DR1*DI2+DI1*DR2
2470 REM Substitution
2480 LET AR2=AR: LET AI2=AI
2490 LET BR2=BR: LET BI2=BI
2500 LET CR2=CR: LET CI2=CI
2510 LET DR2=DR: LET DI2=DI
2520 RETURN
2530 REM Subroutine 2530
2540 LET R=AR2*ZO+BR2+CR2*ZO*ZO+DR2*ZO
2550 LET X=AI2*ZO+BI2+CI2*ZO*ZO+DI2*ZO
2560 GOSUB 2720
2570 LET MP=M: LET VP=V
2580 LET R=AR2*ZO+BR2-CR2*ZO*ZO-DR2*ZO
2590 LET X=AI2*ZO+BI2-CI2*ZO*ZO-DI2*ZO
2600 GOSUB 2720
2610 LET M11=M/MP: LET V11=V-VP
2620 LET R=AR2*DR2-AI2*DI2-BR2*CR2+BI2*CI2
2630 LET X=DI2*AR2+AI2*DR2-BR2*CI2-BI2*CR2
2640 GOSUB 2720
2650 LET M12=2*ZO*M/MP: LET V12=V-VP
2660 LET M21=2*ZO/MP: LET V21=-VP
2670 LET R=-AR2*ZO+BR2-CR2*ZO*ZO+DR2*ZO
2680 LET X=-AI2*ZO+BI2-CI2*ZO*ZO+DI2*ZO
2690 GOSUB 2720
2700 LET M22=M/MP: LET V22=V-VP
2710 RETURN
2720 REM Subroutine 2720
2730 LET M=SQR(R*R+X*X)
2740 IF M=0 THEN LET V=O: RETURN
2750 IF (X/M)>=.999999 THEN LET V=PI/2: RETURN
2760 IF (X/M)<=-.999999 THEN LET V=-PI/2: RETURN
2770 IF ABS (R)<=1E-09 THEN GOTO 2810
2780 IF R<0 THEN GOTO 2800
2790 LET V=ATN (X/R): RETURN
2800 LET V=PI+ATN (X/R): RETURN
2810 IF X=0 AND R>0 THEN LET V=0: RETURN
2820 IF X=0 AND R<0 THEN LET V=PI: RETURN
2830 IF X>0 THEN LET V=PI/2-ATN (R/X): RETURN
2840 LET V=-PI/2+ATN (R/X): RETURN
```

For $k^2 = 2$ and $\theta = 1.5$ radians (see Table 1) the above mentioned responses take the following values: $VSWR_1 = 1.0098$, $VSWR_2 = 1.0656$, $VSWR_3 = 1.0073$, $C_{12} = 4.7761$ dB, $C_{13} = 1.7595$ dB, and $I_{23} = 37.0940$ dB.

Here it deserves noting that for $Z_{01} = Z_{04} = Z_{05} = Z_0$, $Z_{02} = Z_{03} = \sqrt{2}\,Z_0$ and $R = 2Z_0$ the divider under consideration transforms itself into the well known one section equivalent dividing hybrid (see 3,4). *The UPD program is available on disk from the RF Design Software Service. See page 64 for details.* **RF**

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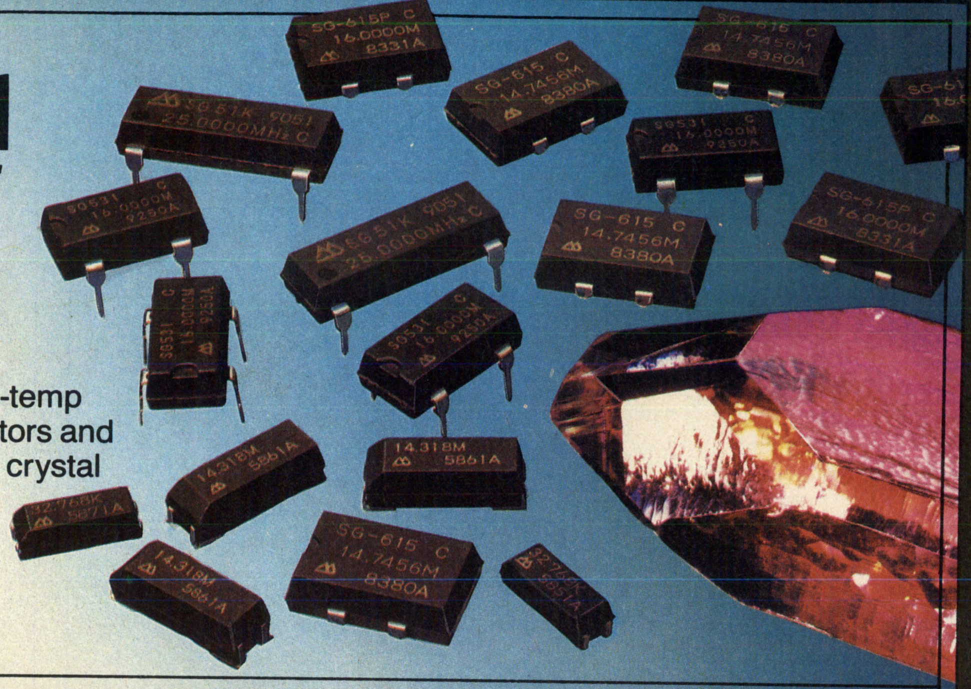

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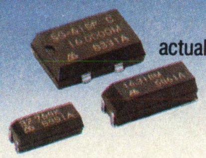

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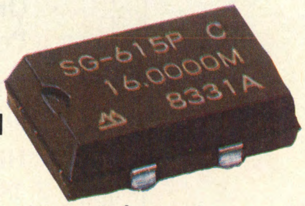

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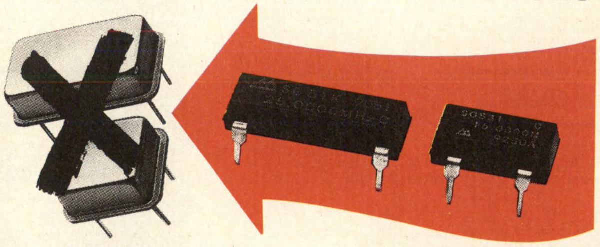

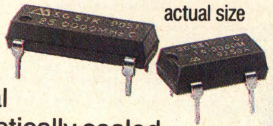

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INFO/CARD 61

Coherent Synthesizer Drives MOTR

By the Engineering Staff
Programmed Test Sources

The AN/MPS-39, a Multiple Object Tracking Radar (MOTR) by General Electric requires precise signals which are switched rapidly to engage various targets in real time. High-purity synthesized frequencies covering bandwidths of hundreds of megahertz are required. These signals must have low spurious and phase-noise content and be capable of being switched rapidly and phase-coherently. Synthesizers employing dividers in their resolution sections will not produce phase-coherent signals when switched at arbitrary times. Figure 1 compares phase-continuous and phase-coherent switching. DDS architecture, which exhibits phase continuity during switching, can be made to behave phase-coherently at the expense of elaborate timing circuitry; but because of spurious considerations, DDS could not fit the bill for this application.

Table 1 lists the specifications for the frequency source required for this MOTR system. The PTS design chosen achieves these specifications with a synthesizer comprised largely of standard modules and a minimum of custom design. This resulted in a highly cost-effective product with predictable MTBF and minimum MTTR owing to the fully modular design. The discussion which follows will focus on aspects germane to the frequency generation only and dispense with ancillary circuitry.

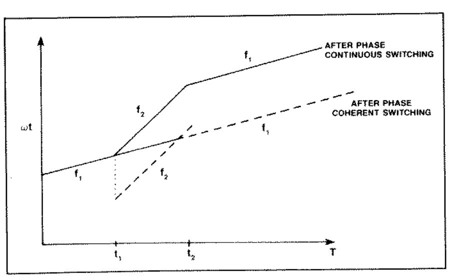

Figure 1. Comparison of phase-continuous and phase-coherent switching.

Figure 2 shows the block diagram of the unit. Twenty-five 10 MHz steps are generated in the "10 MHz Step Section." In this portion a VCO, stepped in 10 MHz increments, operates in a drift-cancelled loop which eliminates the free-running oscillator's frequency drift. As can be seen, the VCO feeds both

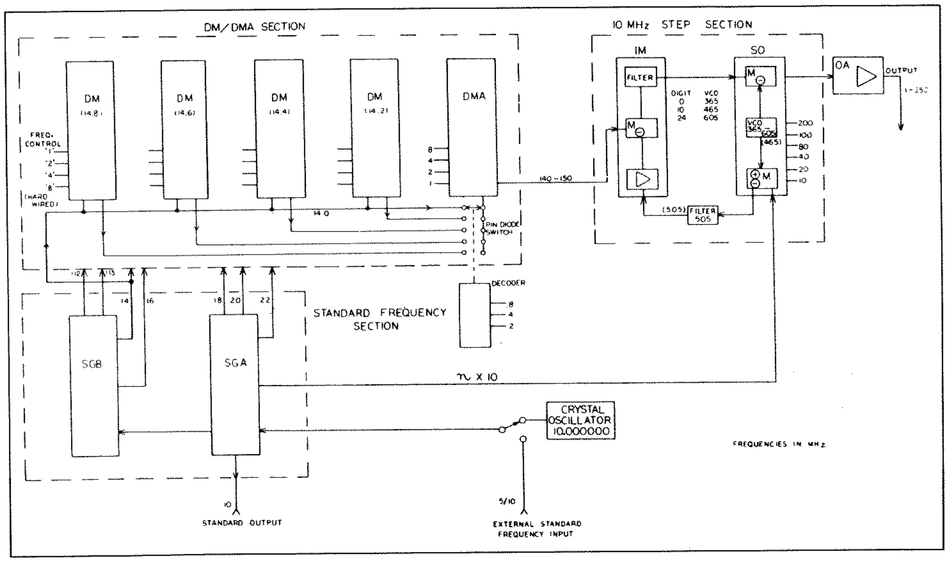

Figure 2. Block diagram, PTS 250 (MOTR).

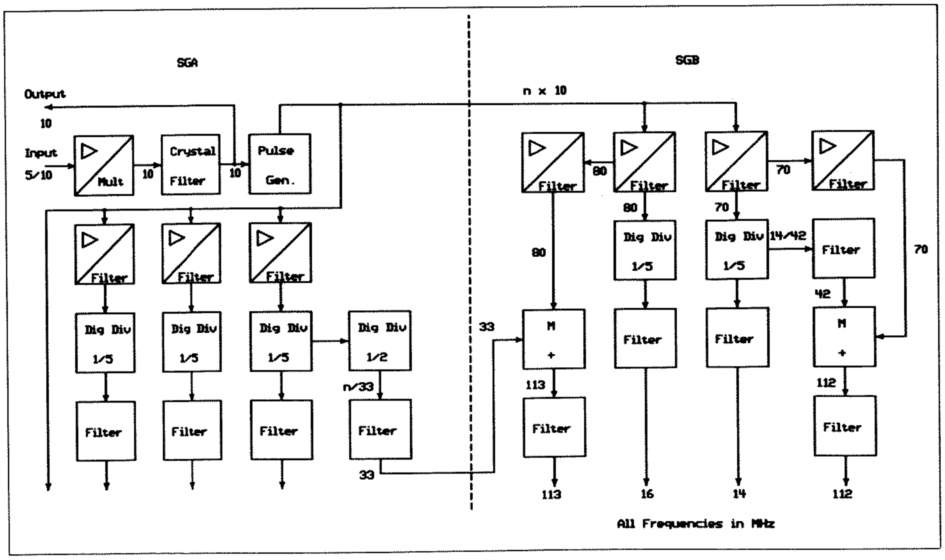

Figure 3. SGA and SGB standard generators.

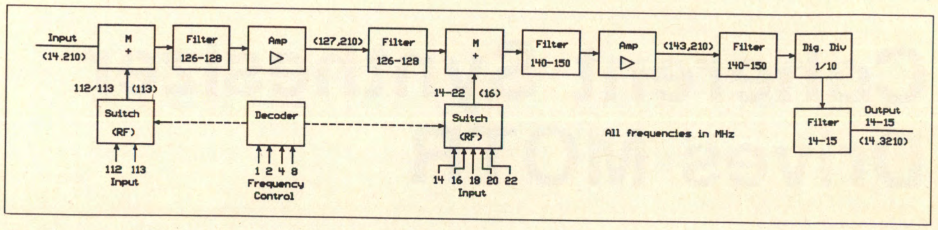

Figure 4. Block diagram of DM module.

inputs to the final mixer, one directly and one after an intermediate mix. Therefore, if the VCO frequency deviates from nominal, both mixer inputs move up or down in frequency together by the same frequency increment. This obviously does not alter the difference of these two frequencies, which is the desired output frequency. The diagram shows some of the VCO frequencies corresponding to certain 10 MHz steps; these should clarify the process. This section essentially selects and filters a 10 MHz line from a pulse generated from the frequency standard. Since all of the 10 MHz lines are continuously generated, frequency switching from one 10 MHz multiple to another occurs phase coherently. Careful temperature compensation of the oscillator and the 505 MHz filter (which is as wide as practical from a point of view of 10 MHz neighboring picket rejection) is essential to minimize phase drift during the observation interval.

The job of fully coherently-switching 1 MHz and 200 kHz steps is accomplished in the DM/DMA sections and by means of auxiliary standard frequencies, derived in the SGA and SGB modules from the same 10 MHz pulse which is used for the 10 MHz steps. These frequencies are then introduced to four DM modules each producing a 200 kHz line and a DMA module which generates the 1 MHz steps.

Steady-State Phase Coherence in Synthesizer Systems

Early frequency synthesizers were not necessarily controlled by a single crystal standard. Adequate frequency stability was obtained by the use of several internal crystal oscillators which contributed to the overall frequency stability of the output. These devices were termed non-coherent. The term phase-coherent as applied to frequency synthesizers describes the relation of frequency-standard to output frequency. If the selected output frequency reproduces accurately the relative frequency stability of the standard, the device is termed coherent.

Despite this, the assumption that all contemporary systems using a single standard (external or internal) are coherent, is not true. Many systems utilizing fractional -n or binary DDS fine-resolution sub-synthesizers are not truly phase coherent, rather they have small but finite reference-to-output errors variously specified.

The SGA and SGB standard generators which operate together, are depicted in the diagram of Figure 3.

An input of 5 or 10 MHz and a level of 0.4V is received from an external or internal frequency standard and fed to the input amplifier/multiplier. This block is followed by a narrow crystal filter. The pulse generator receives its input from this crystal filter and produces a spectrum of n × 10 MHz multiples, which is the basis for all fixed or standard frequencies in the synthesizer. Of these, the five low frequencies of 14, 16, 18, 20 and 22 MHz are produced by division of a specific spectrum-line through an amplifier-filter circuit, a digital divider, and a filter at the output frequency. The two higher standard frequencies of 112 and 113 MHz also use one specific n × 10 spectrum line each. Mixers add derivatives of certain low standard frequencies, as shown, to obtain the final frequencies. Thus, by multiplication, division and addition (arithmetic operations) the SGA and SGB produce seven

Frequency Switching: Phase Continuous or Phase Coherent?

Although the terms phase continuous and phase coherent are sometimes used interchangeably, they actually refer to two distinct properties.

The term phase continuous frequency switching denotes the property that at the switching point the phase of the signal at both the "old" and the "new" frequencies are equal, with no transients or discontinuities. Phase continuous frequency switching is possible in a DDS because of its ability to maintain an accumulated phase value during a frequency switch, and after the next clock pulse begins generating the output signal at the new frequency from the phase value reached by the old frequency.

The term phase coherent as applied to the switching behavior of a signal defines the signal's steady state phase after the switching process is completed. Beginning with two in-phase signals at frequency f_1, assume that one undergoes the switching sequence f_1, f_2, f_1. If after the switching sequence, the two signals are again in-phase, phase coherent switching has occurred. In general, with arbitrary timing the phase transients required for phase-coherence preclude phase continuity.

Figure 1 displays a signal undergoing frequency switching at time t_1 and t_2. With phase-continuous switching (shown by the solid line ______) at time t_2 the signal returns to frequency f_1, but with the phase now offset from that of the original unswitched signal (shown by the dashed line - - -). To obtain a phase coherent switching sequence, in general some phase discontinuity must take place (shown by the dotted line . . . .).

standard frequencies which are completely coherent with the input frequency of 5 or 10 MHz. Three of the five lower frequencies are produced in the SGA.

Since the sections are similar, a description of one - the 20 MHz generator - shall suffice to illustrate the process. Loosely coupled to the spectrum bus (n × 10) by a filter, a transistor amplifies the 100 MHz line, and this signal, after filtering reaches a divider. A divide-by-five is used, and the 20 MHz signal is connected to a tuned circuit. A low impedance output is fed to the 20 MHz bus in the deck for distribution.

A frequency of 33 MHz is needed to produce 113 MHz in the SGB module. This signal is produced by division of 22 MHz by 2. The third harmonic of 11 MHz, after filtering, is fed to the SGB module.

As shown the SGB receives the 10 to 140 MHz spectrum from the SGA module. The low frequencies are produced by the same divide-by-five process which is used in the SGA module: the 70 MHz line is pre-selected, amplified and filtered before it enters the divider; the resulting 14 MHz signal, after filtering, reaches the output.

The same IC divider also supplies a third harmonic of 14 MHz. After amplification this 42 MHz signal serves as injection to the final mixer producing 112 MHz by adding a 70 MHz input.

Although dividers are used in various stages of the standard frequency processing, it should be noted that they are all operating in a C/W fashion, with no switched inputs, thus producing coherent signals once the unit is powered. These frequencies are then introduced to four DM modules each producing a 200 kHz line and a DMA module which generates the 1 MHz steps. The DM modules, ordinarily utilized in series and with a selectable output, are hardwired to produce only one frequency and are operated in parallel. The block diagram of Figure 4 shows the process inside each module. The 14.0 MHz input is up-converted to 126 MHz, and for each of the four modules one of the four bus frequencies 16-22 MHz is added. After filtering, amplification and division by 10, output frequencies of 14.2, 14.4, 14.6 and 14.8 MHz are obtained; (14.0 MHz is directly available from the bus). These 200 kHz raster signals (14.0 to 14.8 MHz) are then selected by pin-diode switches from the external remote frequency control for the 200 kHz steps.

The realization of the 1 MHz steps is similar to the above, except that here no parallel modules are required. The block diagram of the DMA module is identical to the DM diagram without the final division. Since all supply frequencies to this module, namely the 14.2 and 14.8 input from the selected DM module, the 112/113 MHz and the 14 to 22 MHz bus frequencies, are all coherent, frequencies generated in the 140-150 band are also coherent. A bi-quinary selection process used the appropriate supply frequencies to cover the 140-150 band in ten 1 MHz steps, with the actual frequency produced dependent on the remote-frequency control input. *RF*

MULTIPLE OBJECT TRACKING RADAR SYNTHESIZER SPECS

Frequency Range:	1 - 250 MHz
Resolution:	0.2 MHz
Control:	BCD
Switching Time:	6μs
Output Level:	+10dBm (into 50 ohms)
Flatness:±0.5dB	
Spurious Signals:	−68dB
Harmonics:	−30dB
Phase Noise: (at any output frequency)	105dBc/Hz at 100 Hz 115dBc/Hz at 1 KHz 123dBc/Hz at 10 KHz 127dBc/Hz at 100 KHz
Switching Behavior:	Phase Coherent

Table 1. Required frequency source specifications for MOTR Synthesizer.

About the Author

Programmed Test Sources, Inc. has been designing and manufacturing high performance frequency synthesizers for 15 years. The company address is 9 Beaver Brook Road, P.O. Box 517, Littleton, MA 01460, and its telephone number is (508) 486-3008.

A Truly Inexpensive VHF/UHF RF Switch

By Andrew Singer
Sinclair Radio Labs, Inc.

The PIN diode switch is often used in attenuating or switching RF signals. A disadvantage is that the cost can be high. The circuit described here uses common 1N4148 diodes, which makes it attractive even for commercial products.

The insertion loss of an RF switch is often the most critical parameter for the system designer, since this loss may add directly to the noise figure of the

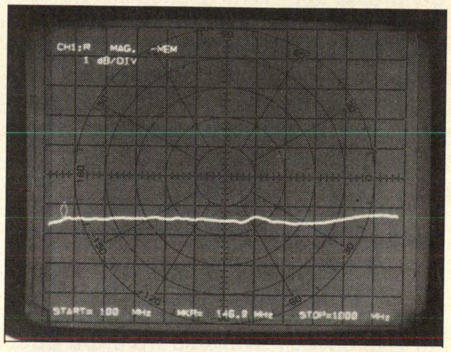

Figure 1. Insertion loss at 100-1000 MHz.

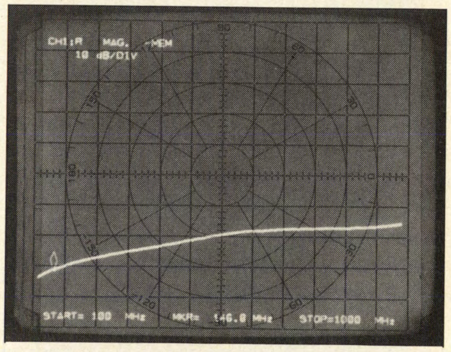

Figure 2. Isolation with no bias voltage applied.

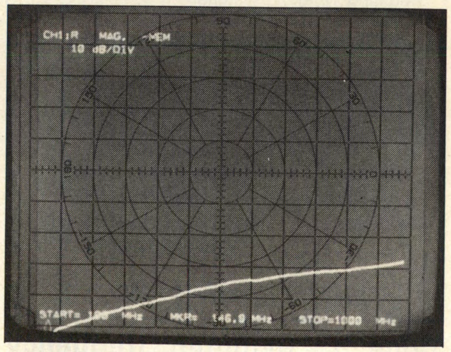

Figure 3. Isolation with -12 VDC bias.

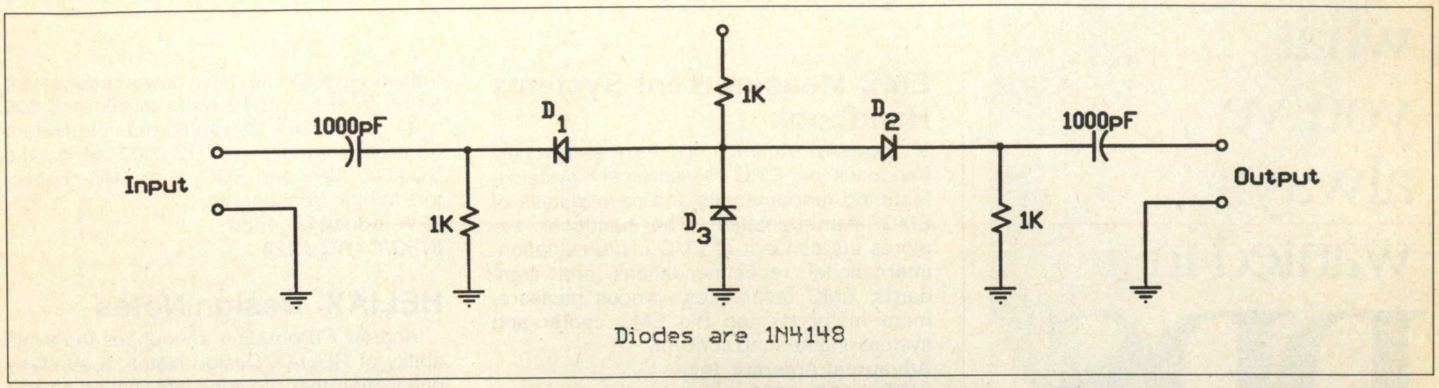

Figure 1. An inexpensive VHF/UHF RF switch.

system. Careful circuit construction can go a long way towards improving insertion loss and isolation. By keeping the diode packages close to the ground plane, the high package inductance can be reduced, which in turn reduces the insertion loss. This also reduces the series coupling capacitance of the package, thus increasing the isolation at higher frequencies (1).

Figure 1 shows the insertion loss from 100 to 1000 MHz to be approximately 1.5 dB. Figure 2 shows the isolation with no bias voltage applied. With a negative bias voltage at A, D1 and D2 are off and D3 will short the signal to ground. This achieves a very high isolation of 30 dB to over 50 dB, as can be seen in Figure 3. Figure 4 shows the circuit. With 12 volts of positive bias at A, the diodes D1 and D2 are forward-biased. The typical packaged series PIN diode would exhibit an isolation of 13 dB to 35 dB in this frequency range with approximately 0.5 dB loss (2).

With a cost of literally pennies per switch, this circuit provides excellent insertion loss and isolation characteristics for VHF/UHF applications. The circuit was used in an electronically steerable antenna, and should prove useful for commercial products where cost is a restricting concern. *RF*

References

1. "Applications of PIN Diodes," Hewlett Packard Application Note 922, pp. 6-7, May 1987.
2. Jack H. Lepoff, "The PIN Diode, Uses and Limitations," RF Expo, Proceedings, 1986.

About the Author

Andrew Singer is a senior applications engineer with Sinclair Radio Labs, Inc. 675 Ensminger Road, Tonawanda, NY 14150. His telephone number is (716) 874-3682.

RF literature

EMC Measurement Systems Handbook

Advantest, America, Inc., is offering a new handbook on EMC measurement systems, featuring fundamentals and case studies of EMC instrumentation. The handbook explores the concept of EMC instrumentation, international recommendations and standards, EMC techniques, various measurement methods, and the EMC center and system engineering concept.
Advantest America, Inc.
INFO/CARD #230

RF/Microwave Connectors

Amphenol Corporation has released distributor catalog DCC-90 detailing coaxial and twinaxial connectors, adapters and accessories. Featured in the catalog are BNC, TNC, Type N, SMA, SMB, Twinax, UHF, and Mini-UHF connector series. Included are cross references for data transmission and military interconnections, and a guide to 32 popular between series adapters.
Amphenol RF/Microwave Operations
INFO/CARD #229

Power Attenuators

A new catalog has been released by JFW Industries detailing new power attenuators rating from 5 and 10 watts to 100 and 300 watts. The 5 and 10 watt attenuators feature frequency ranges of DC-2000 MHz and DC-4000 MHz and BNC, N or TNC connectors at high frequencies.
JFW Industries, Inc.
INFO/CARD #228

HELIAX[R] Design Notes

Andrew Corporation announces the availability of HELIAX Design Notes, a quarterly publication that provides information on HELIAX coaxial cables. Comparisons of HELIAX cables to other cables are included. Other topics that will be covered include shipboard and aircraft applications, RF shielding, and new product specifications.
Andrew Corporation
INFO/CARD #227

Military Products Databook

The 1990 Military Product Databook detailing devices processed in accordance with MIL-STD-883 is now available from Analog Devices, Inc. The databook lists monolithic devices, hybrid devices, and packaging information for all products.
Analog Devices, Inc.
INFO/CARD #226

SPECIAL COURSES AT RFEE

NOVEMBER 12, 13, 14 AND 15

RF expo EAST November 13-15, 1990 MARRIOTT's Orlando World Center Orlando, Florida

Preregistration is required for the following special courses. Please see the registration card on page 85 for fee.

Fundamentals of RF Circuit Design: Part I — Nov. 12

This highly popular course provides an introduction to RF circuit concepts without an intimidating amount of complex mathematics. RF component models are reviewed first, explaining parasitic effects, progressing to resonant circuits, filters and impedance matching principles, comparing analytical and graphical (Smith Chart) techniques. Scattering (s-) parameters, unilateral and bilateral small-signal/low noise amplifier design methods, stability conditions, constant gain circles are illustrated by the use of video-projected interactive CAD. Instructor: Les Besser, president, Besser Associates Inc.

Fundamentals of RF Circuit Design: Part II — Nov. 13

A sequel to Part I, this newly revised course begins with microstrip transmission line applications in RF circuits. Transmission line and transformer type power dividers and combiners, wide-band "multifilar" ferrite-core autotransformers (rod and toroid) are examined under "real life" conditions, considering balance, isolation and impedance transformation. PIN diode switches and attenuators are analyzed by linear circuit simulators. Broadband feedback and high-power amplifiers are reviewed; the effects bias, temperature, parasitics and losses are considered, and an introduction to tolerance analysis is presented. Instructor: Les Besser, president, Besser Associates Inc.

Computer-Aided Filter Design — Nov. 12

This course covers L-C filter design and analysis, with emphasis on principles and solving practical problems using the personal computer. The historical development of filter disign introduces the course, followed by discussions of the modern lowpass prototype, its transformation to highpass, bandpass and bandstop configurations, and the transfer function amplitude and delay characteristics of each. Additional topics include elliptic transfer function filters, determination of required filter order, effects of component and Q parasitics, filters as matching networks, and comments on distributed (transmission line) filters. Instructor: Randy Rhea, Eagleware/Circuit Busters.

Oscillator Design Principles — Nov. 14

First time offered at Expo East. Learn the fundamentals of oscillator design. Historically, oscillator design has been obscured with pages of equations for particular configurations. In this course, basic concepts are applied to design various oscillators using a unified approach. Attendees learn how to evaluate oscillator designs accurately. L-C distributed element, SAW and crystal oscillators are studied. Also considered are output level, starting time, harmonic levels and phase noise performance. Instructor: Randy Rhea, Eagleware/Circuit Busters.

Introduction to Modern CAD Techniques — Nov. 15

This new course offers a complete overview of the various forms of computer-aided design techniques used by RF/MW circuit and system designers. DC, AC time-domain, harmonic-balance and transient analysis principles are reviewed, including statistical and worse-case considerations. Matching network and filter synthesis are compared for both lumped and distributed elements. Circuit optimization, design centering and system analysis concepts are introduced through illustrative examples. The session closes with a discussion of software/hardware combinations to meet various budget considerations. Textbook: Computer-Aided Design of Electronic Circuits. Instructor: Les Besser, president, Besser Associates Inc.

RF EXPO EAST is sponsored by *RF Design* magazine.

RFEE '90 EXHIBITORS

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American Technical Ceramics
Amphenol RF
Amplifier Research
Analogic Corp.
Analog Devices, Inc.
Applied Engineering Products
Applied Specialties
Avantek, Inc.
Aydin Vector Division
Boonton Electronics Corporation
Brennan Associates
Burr-Brown
California Eastern Labs
Ceramic Devices, Inc.
Compac Development Corp.
Compact Software Inc.
Compex Corporation
CTS Corporation - Frequency Control Div.
Direct Conversion Technique
Eagleware/Circuit Busters
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Hewlett-Packard
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ITT Sealectro
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Loral Microwave - FSI Semiconductor
M/A-COM
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Millimeter Wave Technology
Motorola Inc.
Murata Erie North America
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TRM Inc.
Trontech, Inc.
TTE, Incorporated
Vectron Laboratories, Inc.
Voltronics Corporation
Wavetek Corporation
Webb Laboratories

Please see the Technical Preview on the following page and use the registration card on pg. 85.

PROGRAM AT A GLANCE

RF expo EAST
November 13-15, 1990
MARRIOTT's
Orlando World Center
Orlando, Florida

TUESDAY, NOVEMBER 13 — Exhibits Open 11:00 a.m. - 6:00 p.m.

A-1 Receivers I (Tutorial)	8:30-11:30	Design of Receivers for Electronic Warfare	S. R. Vincent, *Raytheon*
A-2 Power Amplifiers	8:30-9:30	VSWR Performance of Transistor RF Power Amplifiers	R. W. Brounley, *Brounley Engineering*
	9:30-10:30	A Novel Technique for Analyzing High-Efficiency Switched-mode Amplifiers	K. Siwiak, *Motorola*
	10:30-11:30	RF Power Transistors for 200-W Multicarrier Cellular Base Station	K. Vennema, *Phillips*
A-3 PLLs and Synthesizers	8:30-9:30	The "Approximation Method" of Frequency Synthesis	A. D. Helfrick, *TIC*
	9:30-10:30	Digitally Derived Swept-frequency Source for Chirped-pulse Radar Altimeters	G. Whitworth, *Johns Hopkins University APL*
	10:30-11:30	Ten-bit 300-MHz DAC for Direct Digital Synthesis	P. Jordan, *Analog Devices*
A-4 Microwaves	8:30-9:30	System Design Considerations for Line-of-Sight Microwave Radio Transmission	G. M. Kizer, *Rockwell* (Dallas)
	9:30-10:30	Phase Shifter Based Upon Reflectively Terminated Multiport Coupler	M. H. Kori, *Centre for Development of Telematics*
	10:30-11:30	Instrumentation for Radar Reflectively Measurement	D. J. Kozakov, C. W. Sirles, D. A. Thompson, and R. S. Banks, *Millimeter Wave Technology*
B-1 Receivers II (Spread Spectrum)	1:30-2:30	Spread Spectrum with Digital Signal Processing	R. J. Zavrel, Jr., *Stanford Telecom*
	2:30-3:30	Receivers for GPS and GLONASS Spread-spectrum Navigation Systems	J. Danaher, *Structured Systems*
	3:30-4:30	Detection and Sorting of Frequency-hopped Signals	J. E. Dunn and S. P. Russell, *Iowa State University*
B-2 MMICS	1:30-2:30	GaAs MMIC Switch Products—Daily Applications	M. D. Smith, *ANZAC*
	2:30-3:30	Silicon MMICs: 35 dB - 35 dBm - $35	J. Walsh, *SGS-Thomson*
	3:30-4:30	Some Design Considerations for L-band Power MMICs	R. Weber, *ISU Microelectronics Research Center*
B-3 Filters	1:30-2:30	Capabilities and Applications of SAW-coupled Resonator Filters	A. Coon, *RF Monolithics*
	2:30-3:30	At Long Last: Modular, Digitally Tuned RF Filters as Easy as Amplifiers and Mixers	E. A. Janning, *Pole-Zero*
	3:30-4:30	Narrow-band SAW Filters for IF Applications	B. Horine and S. Gopani, *Sawtek*
B-4 Antennas and Propagation	1:30-2:30	Large Loop Antennas	R. P. Haviland
	2:30-3:30	VHF Multipath Propagation: Causes and Cures	D. R. Dorsey, *DocSoft*
	3:30-4:30	Radio-frequency Identification Systems for Commercial and Industrial Applications	J. Eagleson, *Allen-Bradley*

WEDNESDAY, NOVEMBER 14 — Exhibits Open 11:00 a.m. - 6:00 p.m.

C-1 Receivers III (Digital)	8:30-9:30	IF Frequency-response Considerations for the Digital Radio Environment	R. Roberts, *Harris*
	9:30-10:30	DSP Demodulation	S. F. Russell, *Iowa State University*
	10:30-11:30	A DSP PSK Modem for SATCOM SCPC Voice/Data	Y. S. Rao, R. Asokan, and K. Reeta, *Centre for Development of Telematics* (India)
C-2 Transmitters	8:30-9:30	Architecture of HF-VHF Radio Transmitters	M. A. Sivers and S. V. Tomashevitch, *Leningrad Electrotechnical Institute of Communications*
	9:30-10:30	Digital-feedback Techniques for a Pulse-width-modulated Power Supply for RF Power Amplifiers	H. Direen, *ETO*
	10:30-11:30	Increasing the Efficiency of SSB Transmitters by Envelope-tracking RF Power Amplifier	L. Voronov, *Leningrad Electrotechnical Institute of Communications*
C-3 SAW Tutorial	8:30-11:30	Surface Acoustic Wave (S.A.W.) Technology	C. A. Erikson, Jr., *Oakmont*
C-4 RF Systems for Research in Particle Physics	8:30-9:30	RF Systems for the Advanced Photon Source	J. F. Bridges, *Argonne National Laboratory*
	9:30-10:30	RF Applications in Particle Accelerators	C. Hovater, *Continuous Electron Beam Accelerator Facility*
	10:30-11:30	A Fully Digital RF-signal Synthesis and Phase Control for Acceleration in COSY	H. Meuth, *Forschungszentrum KFA Julich* (FRG)
D-1 Receivers IV (Applications)	1:30-2:30	Dynamic Evaluation of High-speed, High-resolution D/A Converters	J. Colotti, *CSD/Telephonics*
	2:30-3:30	A High-performance Low-cost TV Demodulator	P. R. M. Correa (BRAZIL)
	3:30-4:30	Digital-signal-processing-based Spectrum-monitoring System for the European Broadcasting Area	I. Novak, *Design Automation, Technical Univ of Budapest*
D-2 Components	1:30-2:30	Impedance-matching Transformers for RF Power Amplifiers	D. N. Haupt, *Erbtec*
	2:30-3:30	Temperature-compensating Attenuators	P. F. Hamlyn, *ANZAC*
	3:30-4:30	Using Current Feedback Amplifiers in High-frequency Active-filter Applications	A. Neves, *PMI*
D-3 Quartz Crystals and Applications	1:30-2:30	Vibrational Sensitivity and Phase Noise in Crystal Oscillators	G. Kurzenknabe, *Piezo Crystal*
	2:30-3:30	Quartz-crystal Filters: A Review of Current Issues	M. D. Howard and R. C. Smythe, *Piezo Technology*
	3:30-4:30	Frequency Correlation of Quartz-crystal Resonators	B. Long, *Piezo Crystal*
D-4 Modulation	1:30-2:30	New Method of Linear Amplitude Modulation	L. Minggang, G. Liangcai, and Z. Suwen, *Wuhan University* (PRC)
	2:30-3:30	4-GHz Multiplied Source for Digital Modulation	D. Balusek, *Rockwell* (Dallas)
	3:30-4:30	Techniques in Voice Compression and Synthesis	P. G. Beaty, *Erbtec*

THURSDAY, NOVEMBER 15 — Exhibits Open 10:00 a.m. - 2:00 p.m.

E-1 Noise Tutorial	8:30-11:30	Noise Fundamentals	F. H. Perkins, Jr., and A. Ward, *RF Monolithics, Avantek*
E-2 CAD and Simulation	8:30-9:30	Spice Modeling and Simulation of an 800-MHz, Class-AB Push-pull Amplifier	R. Y. LaLau, *Mobile Data*
	9:30-10:30	Computer Programs Design and Optimize High-efficiency Switching-mode RF Power Amplifiers	N. O. Sokal and I. Novak, *Design Automation*
	10:30-11:30	A Quasi-linear Determination of UHF Power Device Operation from a Spice-simulated Nonlinear BJT Model	P. E. D'Anna, *MMD*
E-3 PLLs and Synthesizers II	8:30-9:30	Designing with Direct Digital Frequency Synthesizers	F. A. B. Cercas, A. A. Albuquerque, and M. Tomlinson, *Instituto Superior Tecnico* (Portugal)
	9:30-10:30	Direct-digital Waveform Generation Using Advanced Multi-mode Digital Modulation	B-G. Goldberg, *Sciteq*
	10:30-11:30	Optimum PLL Design for Low Phase-noise Performance	S. Goldman, *E-Systems*
E-4 Test and Measurement	8:30-9:30	Design and Development of an RF Data Acquisition System	T. H. Jones and M. A. Belkerdid, *Univ of Central Florida*
	9:30-10:30	Handheld Probing Techniques for RF-PCB and Hybrid-circuit Characterization	Y. D. Kim, *Hewlett-Packard*
	10:30-11:30	Using the VXI Bus for RF Test Equipment	M. Levy, *Racal-Dana*

SMB and SMC Connectors

A catalog of SMB and SMC RF connectors has been released by M/A-COM Omni Spectra. These miniature coaxial connectors can be used in radios, medical instrumentation, cellular radio and industrial measuring and control instrumentation. This line of connectors includes: crimp and clamp attachments for cable, panel mount receptacles, and printed circuit board mount versions.
M/A-COM Omni Spectra, Inc.
INFO/CARD #225

Ferrites for Surface Mount Transformers

A data book from the Special Products Division of Siemens Components, Inc. provides information on the Siemens line of ferrites and accessories for surface mount transformers and sets. The data book details electrical and mechanical information on bobbins for ferrite core shapes and includes information on packaging for automatic assembly.
Siemens Components, Inc.
INFO/CARD #224

Surface Mount Inductors

A four page brochure describing surface mount inductors from 0.1 μH to 1000 μH is available from J. W. Miller (division of Bell Industries). The operating temperature range is −20 to +85 degrees Celsius and current ratings range from 740 MA/0.1 μH through 30 MA/1000 μH.
J. W. Miller Division
INFO/CARD #223

RF Coaxial Connectors and Switches

Kings Electronics has released a catalog on its RF coaxial connectors and switches for Traffic Alert and Collision Avoidance Systems (TCAS). It details plugs, jacks, receptacles, and adapters, as well as cabling procedures.
Kings Electronics Company, Inc.
INFO/CARD #222

Quartz Crystals

Quartz crystals up to 60 MHz and oscillators up to 55 MHz are described in a new catalog from Epson America, Inc. Crystals include cylinder and plastic package SMD types. Crystal oscillators include SIP, full and half size DIP, standard and small SMD, and programmable oscillators.
Epson America, Inc.
INFO/CARD #221

Advertising Index

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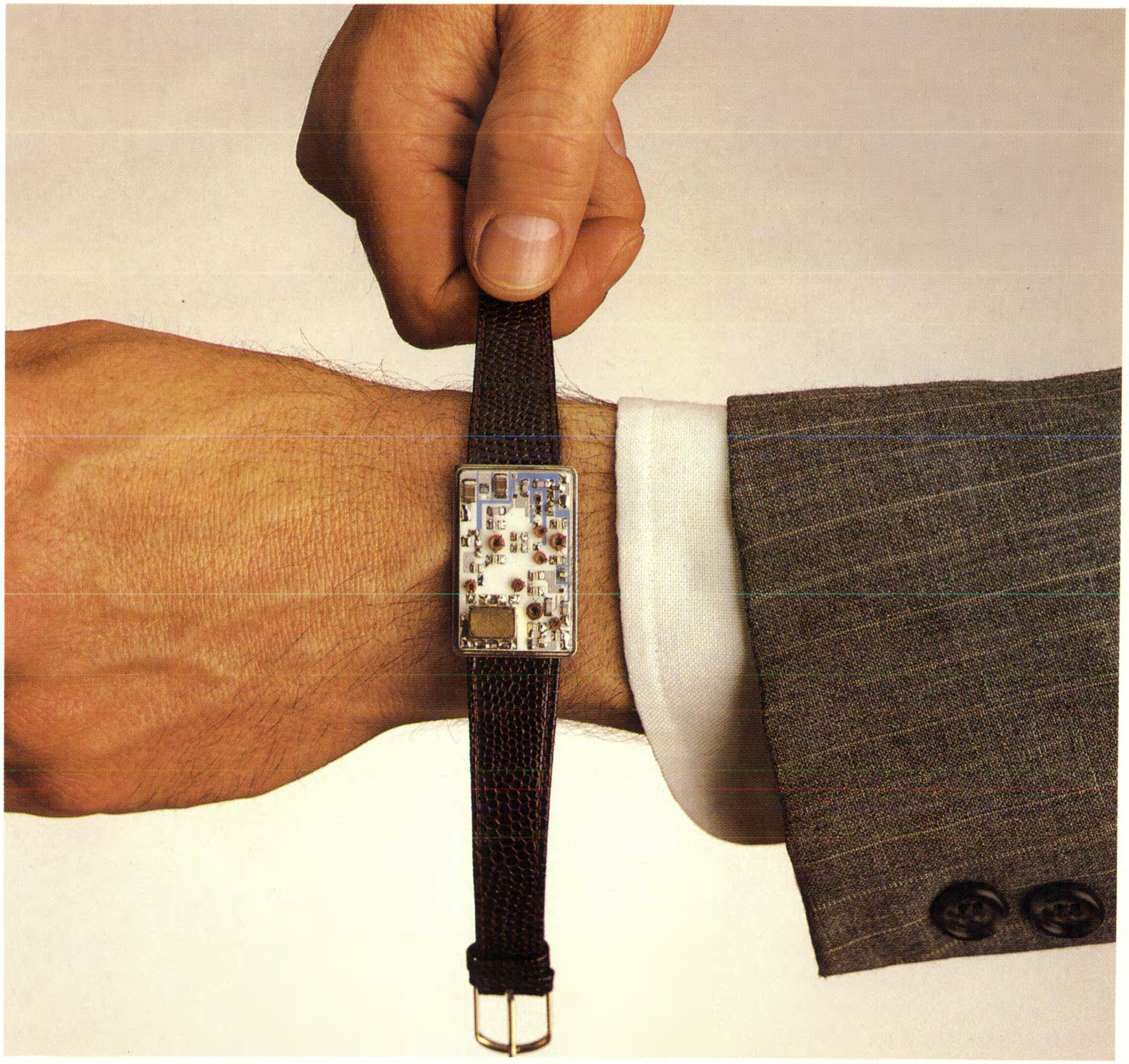

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