RF design

engineering principles and practices

September 1991

Cover Story
New dB-Linear AGC Amplifier

Featured Technology
Piezoelectric Devices

Product Report
High Performance Software

Frequency Response Testing Made Affordable

A-7550 Spectrum Analyzer with Built-in Tracking Generator

The IFR A-7550 Spectrum Analyzer with its optional built-in Tracking Generator may be all the test equipment you need to test the frequency response of any frequency selective device between 10 kHz and 1000 MHz. For higher frequency devices, the A-8000 Spectrum Analyzer with its optional built-in Tracking Generator can characterize frequency responses up to 2600 MHz.

With either analyzer you get a rugged, portable instrument that is equally at home in the field, on the manufacturing floor, or in the laboratory.

Other standard features of both the A-7550 and A-8000 include a synthesized RF system, +30 dBm to −120 dBm amplitude measurement range, 1 kHz per division frequency span, and 300 Hz resolution bandwidth. These features give the A-7550 and the A-8000 superior amplitude and frequency measurement capability previously unavailable on spectrum analyzers in this price range.

In addition to the Tracking Generator, other available options—such as an Internal Rechargeable Battery Pack, AM/FM/SSB Receiver, RS-232 or IEEE-488 Interfaces, and Quasi-Peak Detector—allow the A-7550 and A-8000 to be custom configured to solve many other RF testing needs.

For more information or a demonstration, contact your local IFR distributor or representative, or contact IFR directly at 316/522-4981.

IFR SYSTEMS, INC.

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INFO/CARD 2

$t_{Rise/Fall} = 0.0000000007$

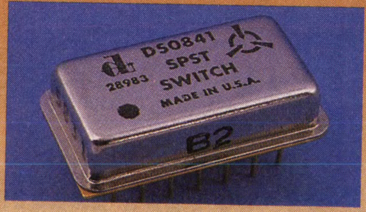
$t_{Rise/Fall} = 0.7x10^{-9}$ seconds

$t_{Rise/Fall} = 0.7$ nanoseconds

$t_{Rise/Fall} = 700$ picoseconds

FAST BY ANY MEASURE!

DAICO Introduces The 700 Picosecond Ultra-Fast Switch



The new DAICO DSO841 offers incredible switching speed. It's much faster than a Schottky diode switch drawing less than 2mA. In fact, Schottky diode switches can draw as much as 100-200mA.

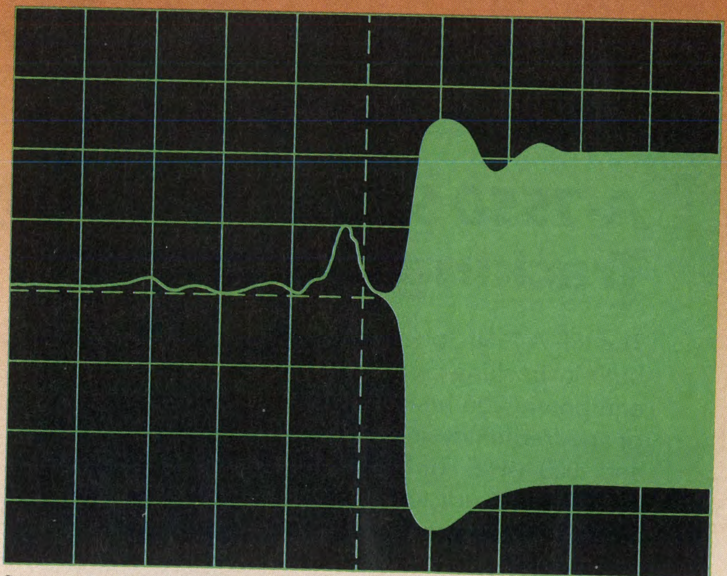
RF envelope rise time (10%/90% RF) for the DSO841 is typically less than 700 psec with 7 nsec switching speed (50% TTL to 90% RF).

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 = 480.00 psec

Call DAICO today at (213) 631-1143 for more information on this new ultra fast switch and our complete line of RF Attenuators, Switches and Phase Shifters.

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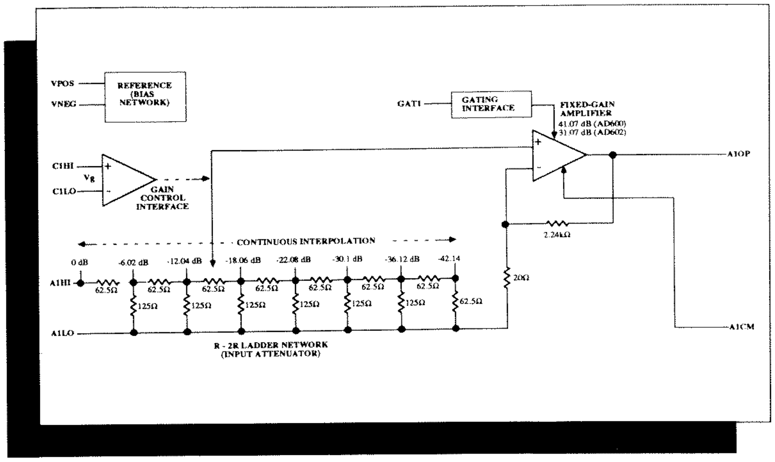
September 1991



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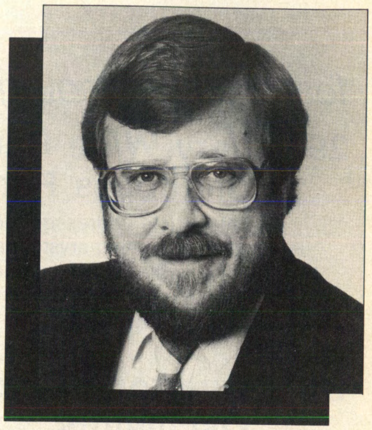
Changes in the RF Business



By Gary A. Breed
Editor

It has been fascinating to watch the RF industry react to recent dramatic shifts in the economy. Seeing how companies cope in tough times is highly educational. Trying to figure out why some firms are busy and very profitable is even more interesting. Here are some observations:

A few companies are definitely in poor shape solely because of the economy, mainly due to military cutbacks. Some companies with a military product emphasis are simply not able to make a rapid shift to commercial applications. The high performance, highly documented MIL process is entirely different than making commercial products in large quantities, with looser specifications.

It is hard to blame the management of the these firms for short-sightedness. After all, even in the worst of times, military hardware is still the largest marketplace for RF technology, especially the performance-oriented technology in which the U.S. excels. But it has been impossible to predict which specific programs would be cut back or eliminated. As a result, a few companies have been squeezed out of this market.

Most companies are suffering from the sluggishness of the overall economy. Those that are remaining competitive and financially stable are working harder and working smarter to do so. New product development has noticeably increased, to broaden a company's base of potential customers, or as product redesign to lower production costs. Unfortunately, some companies are not in good enough financial condition to support new product development when their sales slow down and end up broke, merged, or sold.

Despite difficult times, success stories are fairly easy to find. Companies enjoying good times fall into two categories: those serving the right market at the right time; and those developing unique technologies. The latter of these has been around since the Industrial Revolution. Technology changes rapidly, new ideas abound, and entrepreneurs who develop them can succeed wildly or fail miserably. Those lucky (or brilliant) enough to create widely used products are the success stories we read about. The majority either fail outright or have only marginal success and end up selling out without realizing their dreams.

Being in the right place at the right time with the right product is every businessman's wish. If I knew what it takes to make the right choice, I'd already be rich! But, there are some things that I have observed in companies who enjoy this kind of success:

First, they work hard to do thorough market research and engineering development. Next, they are led by men and women who have clear goals and methods in mind, and stay with them. Finally, they all seem to have a "sixth sense," an intuition that is combined with proven business methods to show them how to proceed.

Note that all of these are human attributes — hard work, vision, dedication, and a little luck. Technology alone is not enough. It is the ability to apply technology in an appropriate manner, and deliver it to the customers who need it, that distinguishes a company which succeeds from those that are struggling.

NEW FROM NOISE COM

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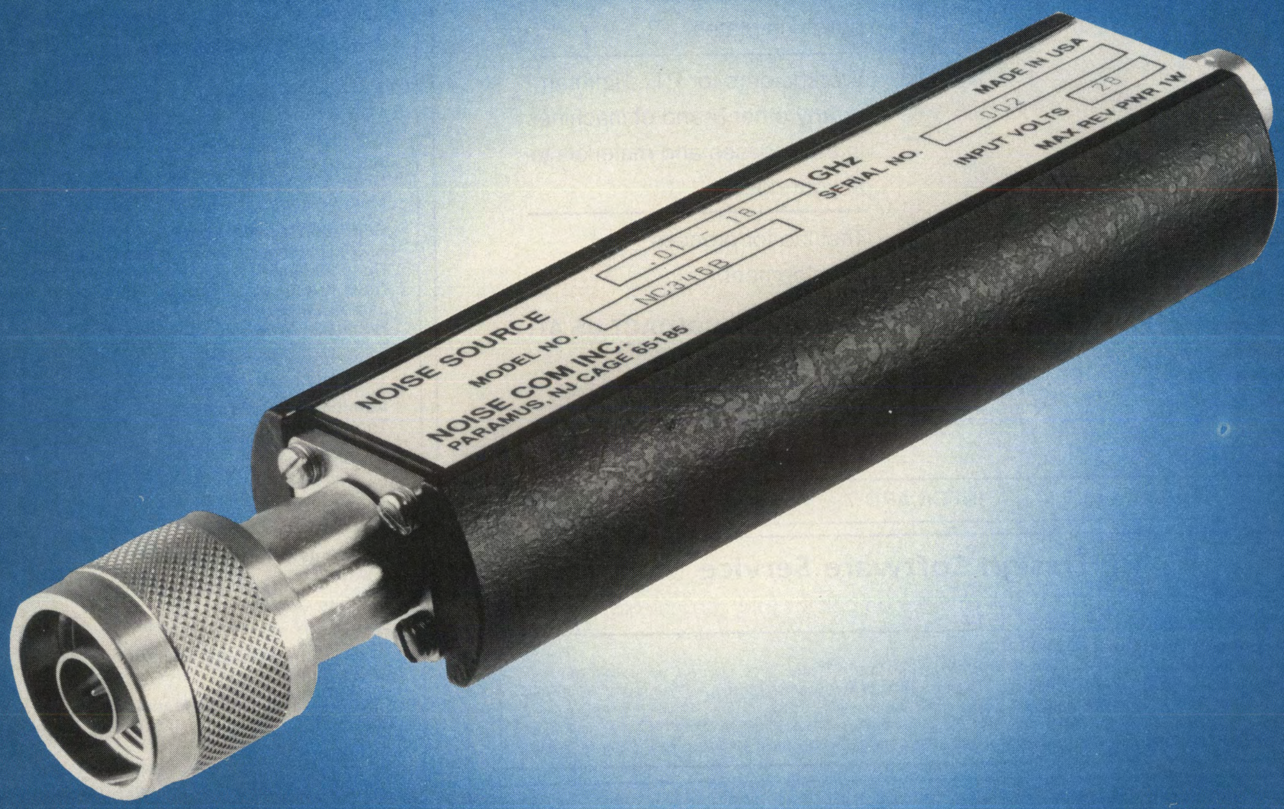

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- ❑ In stock for immediate delivery
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Noise Com designed the NC 346 calibrated noise sources exclusively for noise figure measurement. With two choices of frequency range, three choices of noise output and four choices of connectors, there's an NC 346 noise source for every popular noise measurement system. And each NC 346 comes with a feature found nowhere else: Noise Com's hard-earned reputation for fast delivery and customer support.

For more information about NC 346 noise sources or any of Noise Com's products, call Gary Simonyan at (201) 261-8797.

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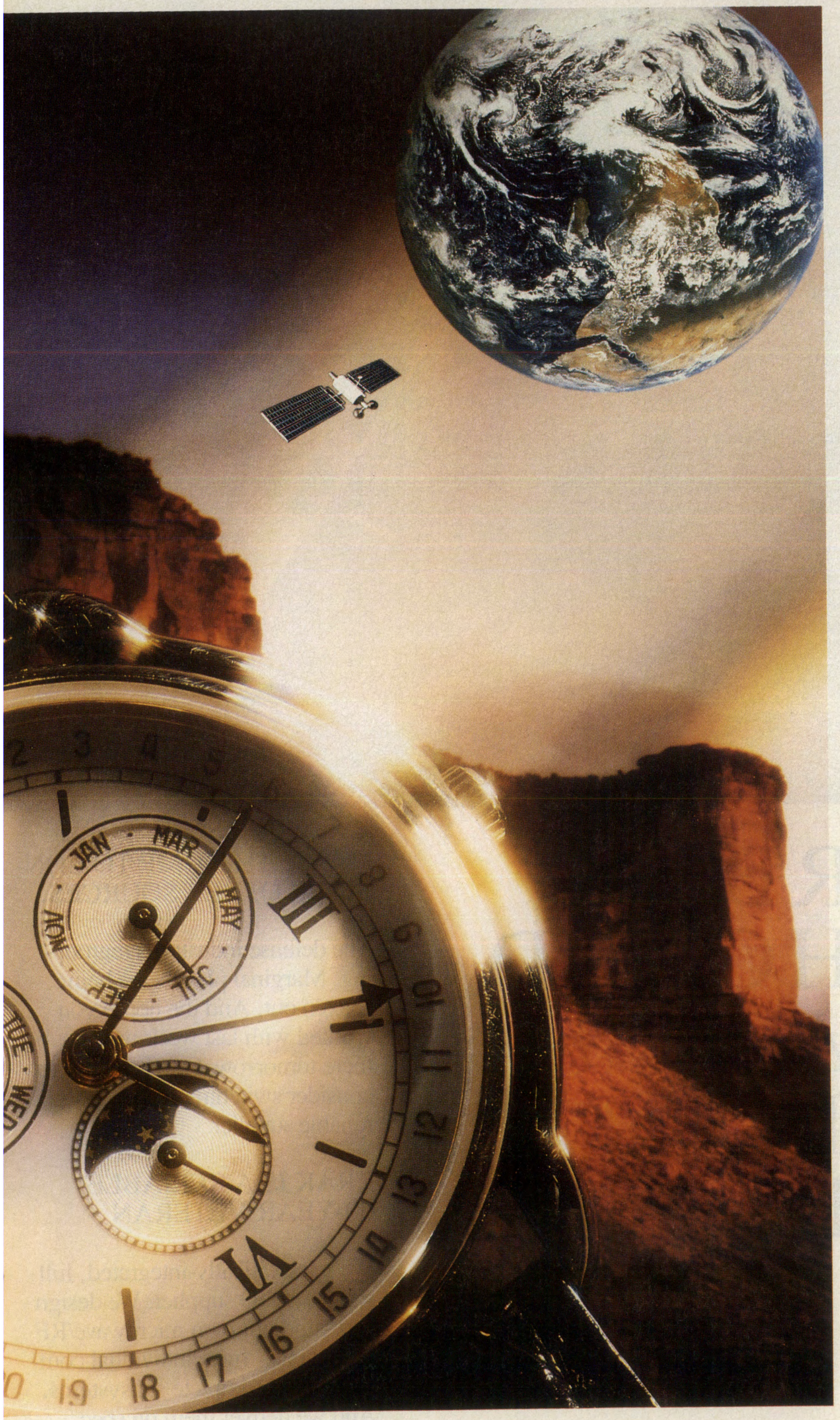
For those who dream of suspending time in the rush to the communications market.



HP helps speed up design cycles by verifying models with real-life signals.

In communication systems labs today, the focus is on improving time-to-market. Comdisco CAE software tools help by creating complex signals to verify models at the design simulation stage. Now, HP's 11755A takes verification a crucial step further, by linking CAE modeling with real-life test equipment to quickly and accurately test brassboards after simulation.

The HP 11755A RF simulator WorkSystem driver provides the link between Comdisco and the HP Vector Arbitrary Waveform Synthesizer. It lets you use real-world signals (10 to 3000 MHz with precision impairments) in the lab to test designs at the earliest possible stage. You get the test results you need to uncover deficiencies and modify designs fast. To speed up design iterations, just download your software signal formats and link testing with CAE software.

So, if you've ever wished you could suspend time in your communications systems design cycle, call your local HP sales office. We'll send you more information on how the HP 11755A can help make your dream come true.

There is a better way.

HEWLETT PACKARD

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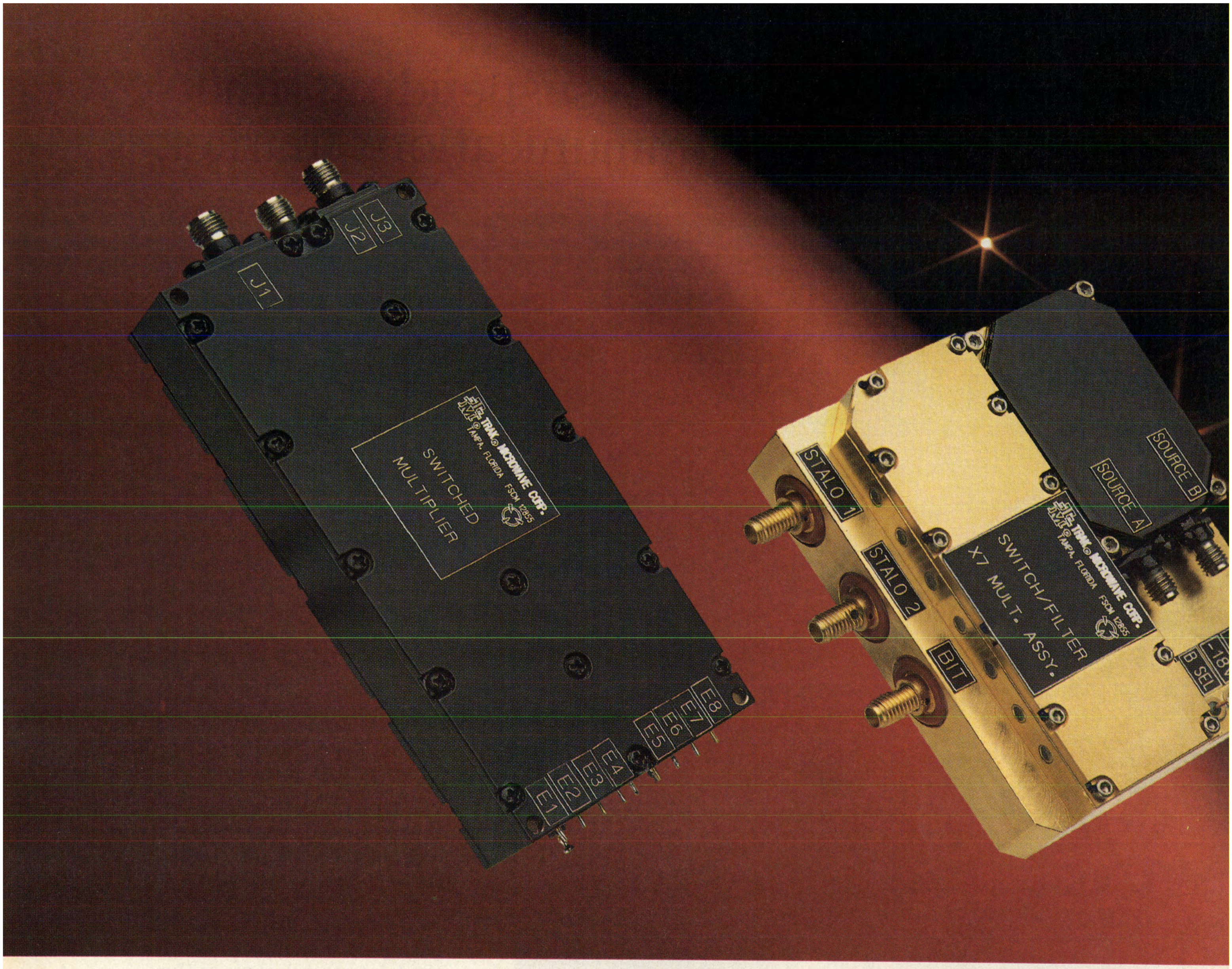

YOUR BEST SWITCH COULD BE MADE BY PICKING UP THE TELEPHONE

NOW'S THE TIME TO CONSIDER OTHER SOURCES.

The defense industry is regrouping. Marginal players are being shaken out. And the house you worked with last week may not be there tomorrow. Good reasons to consider switching to a steady, reliable supplier.

TRAK IS A RELIABLE, 30-YEAR VETERAN

We're a vertically-integrated, full-service, DoD supplier. We design and build our own microwave/RF products — from individual components to complete subsystems. And we deliver them on spec, on time, and on budget.

CONSIDER SWITCHING TO TRAK PRODUCTS

Like our switched filter banks, offering unusually high performance in a small package. Or our switched multipliers, featuring rugged microstrip circuitry in a small, low-weight envelope. And if you don't find what you want in our catalogs, call us anyway — we've got products in mind we haven't built yet, just waiting for someone like you to need them.

BUT WHATEVER YOU'RE WORKING ON, BE SURE OF YOUR SUPPLIER. THAT'S WHY IN TIMES LIKE THESE, IT'S TIME TO THINK TRAK.

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

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TOTAL QUALITY LEADERSHIP

INFO/CARD 9

RF letters

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

Power Transistor Comments

Editor:

I read with interest your report "RF Transistors — Staying on Top of Technology." The rapidly changing technology and vast number of applications warrant far more space than this single page article.

In particular, I would like to take exception to the comment that "Static induction transistors have come and gone from the headlines." The Microwave Technology variation of this device, the SST^{tm} or "solid state triode," is gaining rapid acceptance in the RF power marketplace. In fact, it was discussed as part of a well-written review of military applications of power transistors by David Hughes in your June issue. These high power devices will start appearing in sockets requiring the improved linearity, wide gain control range, efficiency and ruggedness that are features of the SST.

So put the SST back in the headlines, and visit us at RF Expo East to find out more about them.

Allen E. Rosenzweig
V.P., Marketing and Sales
Microwave Technology

Editor:

In your "RF Product Report" in the July issue, you stated that Motorola leads the market in RF power transistors at the lower end of the radio spectrum. I would like to make the public aware that we are also the leader in pulsed L-band power transistors. This has been developed, over the past two years.

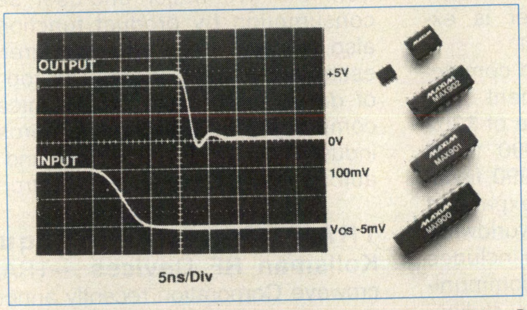
This fact is unknown at this time because we don't have these products in our RF catalog. We currently have 19 new products that range from 5 to 500 watts minimum at duty cycles from 1 to 25 percent, and we are scheduled to have four new products introduced in the third quarter.

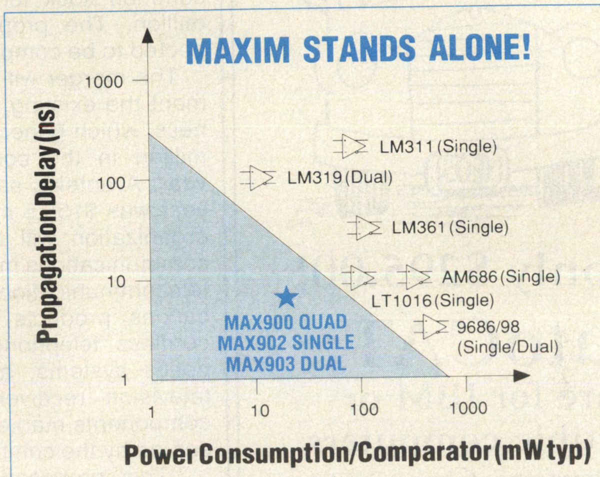
Evaluating current and new geometries for customer specials, to replace devices no longer available, has been the real bottleneck in getting products introduced. Little time is left to fulfill the requirements for new product introduction.

Beth Longbrake
Applications Engineer
Motorola, Inc.

WORLD'S LOWEST POWER HIGH SPEED COMPARATOR-18mW, 8ns!

Maxim now offers you a selection of the fastest low-power single, dual and quad TTL comparators. For example, the new MAX903 can deliver an 8ns response time, while drawing only 3.6mA (18mW) per comparator from a +5V supply (enabling signals in excess of 100MHz to be processed). Although other comparators may operate from a single supply, the MAX900 series are the only high-speed comparators with an input voltage range that extends down to ground. In many applications this eliminates the negative supply, saving board space, power consumption and cost!



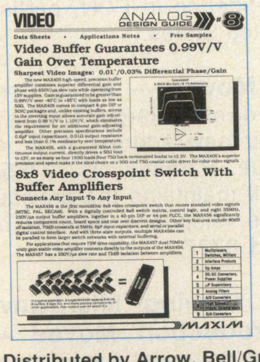


- **Separate Analog and Digital Supplies Allow:**
  Single +5V to +10V Analog Supply
  Split ±5V for Bipolar Operation
- **Input Range Includes Ground and Eliminates Need for Negative Supply**

Select A High Speed Comparator For Your Low Power Application

Device	# Comps	Prop Delay (ns)		Power Consumption (mW)	Single +5V Operation	Input Voltage (Single +5V Supply)	TTL Outputs	Price†
		Typ	Max					
MAX900	4	8	10	70	YES	-100mV to +2.5V	Single Ended	$7.01
MAX901	4	8	10	70	YES	-100mV to +2.5V	Single Ended	$5.98
MAX902	2	8	10	35	YES	-100mV to +2.5V	Single Ended	$4.01
MAX903	1	8	10	18	YES	-100mV to +2.5V	Single Ended	$3.15
MAX912[1]	2	8	10	40	YES	-100mV to +2.5V	Complementary	$4.00
MAX913[1]	1	8	10	25	YES	-100mV to +2.5V	Complementary	$3.13
LT1016[1]	1	10	14	125	YES	+1.25V to +3.5V	Complementary	$3.13

1. Upcoming new products—available after March, 1992



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† 1000-up FOB USA, suggested resale

Hewlett-Packard to Acquire Avantek

Hewlett-Packard Company and Avantek Inc. have announced an agreement for H-P to acquire Avantek, a manufacturer of RF and microwave semiconductors, components and assemblies. Terms of the proposal call for H-P to pay $4.60 per share for each of the approximately 18 million outstanding shares of Avantek common stock, for a total of about $82.8 million. The proposed merger is expected to be completed this fall.

The merger will expand and complement the existing H-P component business, which generated revenue of $302 million in the company's 1990 fiscal year. Avantek's sales in its 1990 fiscal year was $155.5 million. The expanded organization will target the worldwide communications market, which includes telecommunications and data communications products like cellular radios, cordless telephones, high-speed fiber optics systems, and direct broadcast television receivers. Defense-oriented components markets will continue to be served by the combined organization.

"This proposed merger will speed products to the market, streamline service to our mutual customer base and allow us to offer more world-class components faster to customers around the world," said William F. Craven, H-P Vice President and General Manager of the Components Group.

Avantek will become part of H-P's Components Group, headquartered in San Jose, California, with plants in Malaysia and Singapore as well as San Jose. Avantek plants are located in Santa Clara, Newark, Milpitas and Folsom, California and in Frimley, England. Some reduction in Avantek employment is likely as overlapping positions are identified, but no estimates have been given.

E. Oran Brigham, Avantek CEO, who announced in May 1991 that he planned to retire before the end of this year, will leave the company after the merger is completed.

The proposed merger is subject to approval from a majority of Avantek shareholders and completion of antitrust reviews by the Federal Government.

New Study on Hybrid Microelectronics — A new study recently issued by Frost & Sullivan International shows "The U.S. Market for Hybrid Microelectronics." According to the report, the 1989-91 period was a rough one that included shrinkage from 1988 levels, principally due to cutbacks in military spending, on which the market depended heavily. However, the emergence of high-end multi-chip modules in particular will spur hybrids to significant gains from what was a 1990 demand of $6.1 billion. To chart what lies ahead, the report breaks down what is happening according to manufacturing technologies for hybrids, demand from the five major end-user sectors, consumption by product function, and also profiles major manufacturers and estimates market share. The top areas of demand will be in communications, computers and peripherals, aerospace, industrial and instrumentation, and finally automotive and consumer electronics.

TRAK Microwave Acquires Kollsman RF Devices —TRAK Microwave Corporation recently announced the acquisition of Kollsman RF Devices, expanding their product line for commercial applications. TRAK moved Kollsman's operations from St. Petersburg, Florida to their plant in Tampa. Details of the acquisition were not released.

Millicom to Sell Microtel, Change Focus —Millicom Inc. has announced its withdrawal from Microtel Communications Limited, in order to focus its resources on domestic personal communications network development. Microtel will be sold to Hutchinson Telecom Ltd. for approximately $7.4 million. According the J. Shelby Bryan, the Chairman and CEO of Millicom, "The decision to leave Microtel is not a reflection on the prospects of the commercial success of PCN in the United Kingdom, rather it is a reaffirmation of Millicom's confidence in PCN, particularly in the United States."

Noise Com Moves — Noise Com, Inc. has moved into a larger facility in Paramus. Their new address is E. 49 Midland Avenue, Paramus, NJ 07652. Noise Com still maintains the previous facility and the telephone number remains (201) 261-8797.

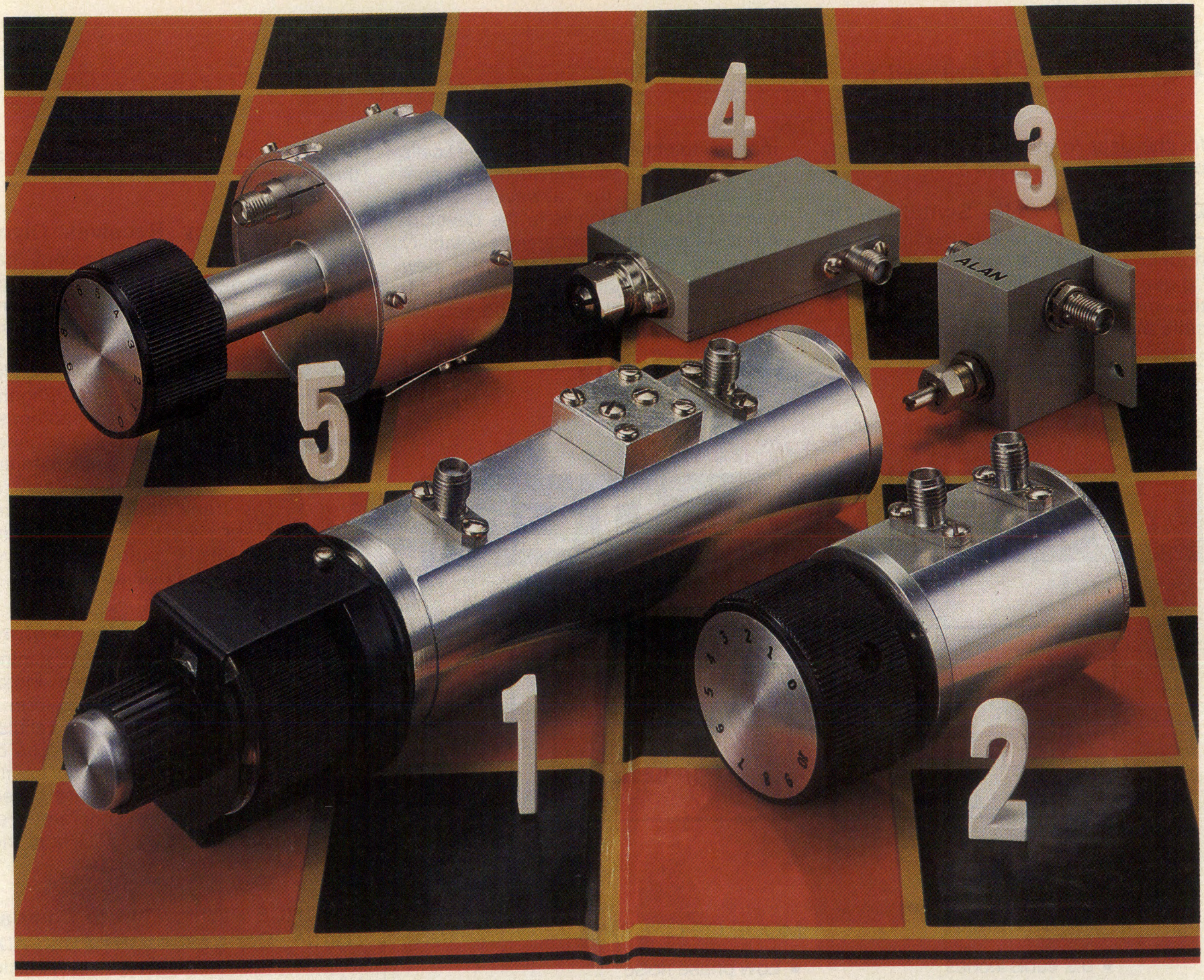
Emerson & Cuming Microwave Absorber Production Moved — Emerson & Cuming recently announced that it will consolidate microwave products fabrication and production at its Canton, Massachusetts facility. The plant located in Gardena, California will be closed.

Micro Linear To Build New Facility — Micro Linear Corporation has started construction on an advanced fabrication facility to manufacture mixed-signal and analog integrated circuits. The new facility will be able to accommodate BiCMOS, high-density CMOS, and high-speed bipolar processes. Completion is expected in Spring of 1992 with prototype chips to be produced from the facility in the first quarter of 1992 and full production expected in the middle of 1992.

AIM and FACT Relocate to Larger Facilities —The Automatic Identification Manufacturers, Inc. (AIM USA), AIM International, and the Federation of Automated Coding Technologies (FACT) announced that they have relocated to larger space. The new address is: AIM USA, 634 Alpha Drive, Pittsburgh, PA 15238. The telephone and FAX number remain the same.

AEL Awarded Option on Contract — AEL Defense Corp has been awarded a $2.5 million option on the Type-18 Periscope Mounted Automatic Direction Finding system, contract by the Naval Sea Systems Command, Department of the Navy. Each system includes such equipment as antenna subsystems, radio frequency receiving devices, a digital processor and a display. With the award of this option, the values of the existing contract exceeds $11.7 million.

ARDIS to Provide NCR Wireless Connectivity —NCR recently announced the inclusion of an integrated Motorola radio packet modem in their new 3125 notepad computer to provide connectivity to the ARDIS nationwide data radio network. This is the first notepad product with an integrated radio packet modem. With ARDIS there are no landline or cellular modem connections


ALAN
1
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5

for the user to initiate with connectivity being transparent to the user.

Digital-Ready Cells Switched On — Bell Atlantic Mobile Systems announced that it has turned on the first "digital ready" cellular transmitters in the mid-Atlantic region. While digital calls are already processed digitally by mobile telephone switches, the calls themselves are still transmitted in analog form over the airwaves. Digital cellular transmission is in the development stage, with commercial availability expected in about a year. The new hardware is capable of handling both analog and digital signals. The network will be able to support dual-mode and full-digital cellular phones as soon as they become available.

RF Technology Receives Olympics Order — RF Technology has received an order from Spanish TV for microwave systems to be used during the 1992 Olympic games. More than 40 systems were ordered including 14 ultra-portable transmitters to be used for wireless camera applications during the games. The portable systems comprise the RF-223 high RF output frequency agile transmitters with complementary RF-200D low noise dual-conversion receivers. The transmitters feature integrated AC or 12VDC power supply, switch selectable RF output power level and wideband transmission capability.

US West Joins Cellular Venture in Japan — US West, Inc. is joining with DDI Corporation, Nissan Motor Company and other partners to form a new company to provide digital cellular telephone service in the Kanto area of Japan. US West will be a minority equity shareholder in the new venture called TU-KA Cellular, which will operate wireless communications at 1.5 GHz in the region.

ADVANCE Partnership to Combat Traffic Congestion — The ADVANCE (Advanced Driver and Vehicle Advisory Navigation Concept) will be tested over the next five years in the Chicago, Illinois area. The project will be the largest field operational test of the Intelligent Vehicle Highway System (IVHS). ADVANCE is designed to provide the motorist with the best route to his or her destination based on current traffic conditions. A fleet of up to 5,000 privately and commercially owned vehicles will be equipped with on-board navigation and route guidance systems. Global Positioning System satellite receivers will help the vehicle's guidance system identify its precise location. Motorola will install and maintain its navigation and route guidance system equipment in the fleet. The company will also provide the RF data communications system for the transmission of data between the test vehicles and the Traffic Information Center. The project will be funded by the U.S. Department of Transportation's Federal Highway Administration, the Illinois Department of Transportation and the remainder by the private sector including the universities and Motorola.

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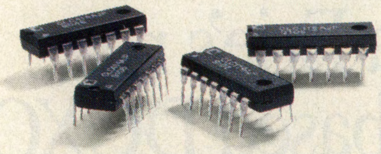
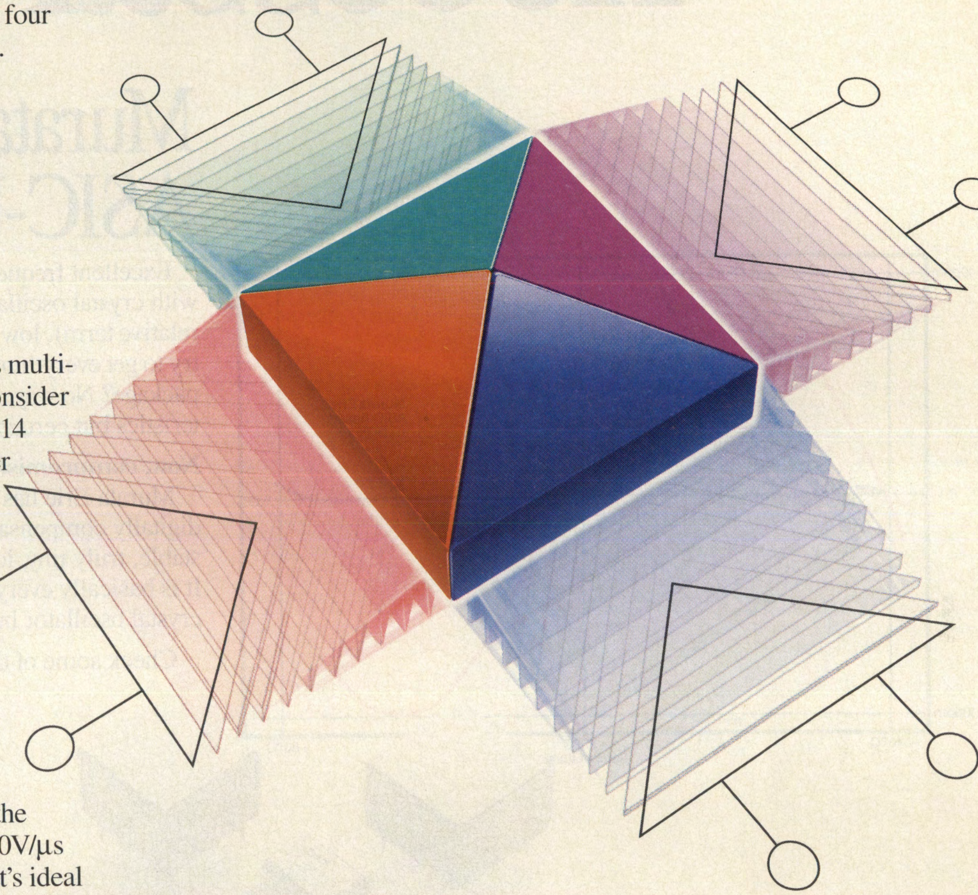
Multi-channel designs have you in a corner? Want to cut cost and space? Then take a look at Comlinear's four new high-speed quad amplifiers.





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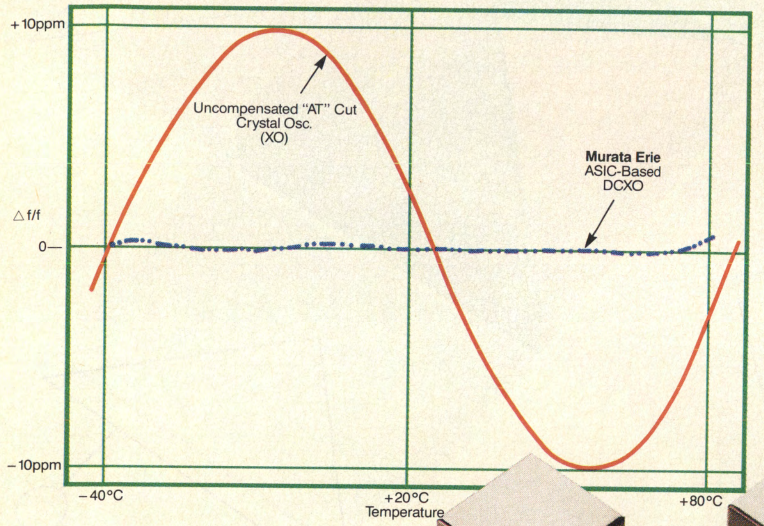

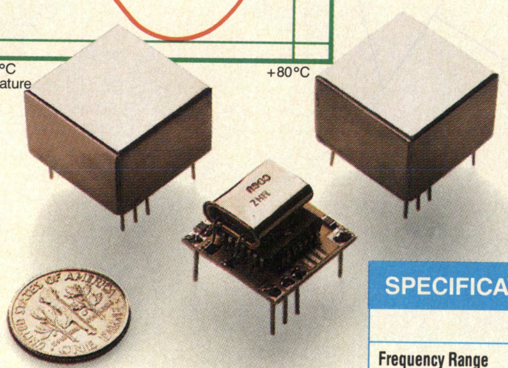

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Now, compromise no more.

Murata Erie has the solution…our new ASIC-based, digitally compensated crystal oscillator is exceptionally stable, truly tiny, low in cost and extremely flexible. It is basically everything that you always wanted in a crystal oscillator but always had to compromise to get.

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- To ±0.2ppm stability
- Less than 5.0 cubic centimeters volume
- 50 mW max power
- Built-in interface, aging compensation, reconfiguration capability and software
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SPECIFICATIONS

	DC2200 AT	DC2210 AH
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Stability over −40°C to +85°C	±0.2 ppm	±1 ppm
Output	TTL	"HC" CMOS
Power Dissipation	+5V @ 10 mA	+5V @ 15 mA
Short Term Stability	1×10E−10 @ T=1 sec.	1×10E−9 @ T=1 sec.
Package Size	2″ × 2″ × 0.5″, PC PINS	0.79″ × 0.79″ × 0.45″, PC PINS

Find out more about how you can take the compromise out of your crystal oscillator specifications. Write for complete details on Murata Erie's new ASIC-based DCXO or call 1-800-831-9172.

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INFO/CARD 20

IBM Makes First High-Temperature Superconducting Magnetic Gradiometer — The first high-temperature superconducting magnetic gradiometer to be operated at liquid nitrogen temperature has been reported by researchers at the IBM T.J. Watson Research Center. A gradiometer is a magnetic measuring device, which, like a magnetometer, uses the most magnetically sensitive detector known, a superconducting quantum interference device, or SQUID. A gradiometer, unlike a magnetometer does not need to be magnetically shielded from external fields like those from earth and from nearby electrical equipment and electronic instrumentation. Fabrication of the gradiometers relies on a unique wet-etching process. The technique depends on a highly selective etchant solution that etches patterns in the thin films of material used to insulate layers of superconductor and stops when the superconductor is reached.

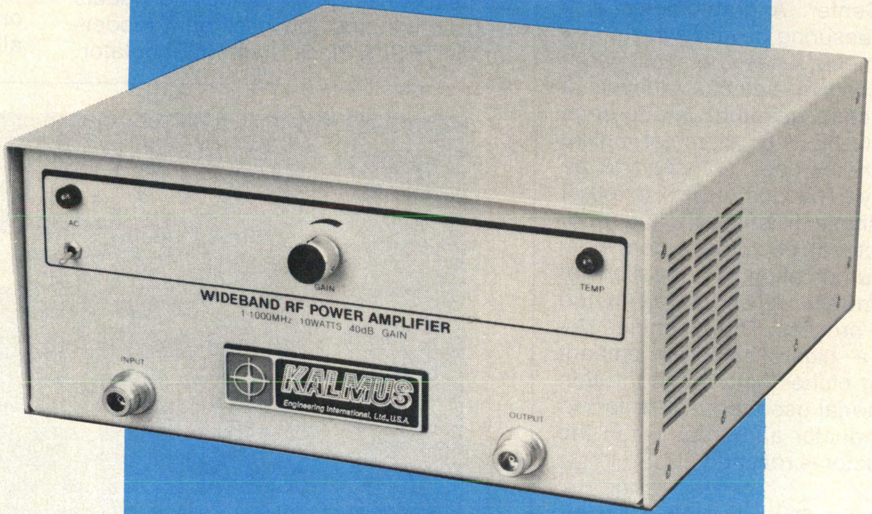
Digital Audio Broadcasting Seminar to be Held —The European Broadcasting Union (EBU), in cooperation with the NAB, will hold an international symposium on digital audio broadcasting in Montreux immediately preceding the NAB Radio Montreux show in June 1992. The symposium will evaluate the situation for DAB after the World Administrative Radio Conference in March 1992 and assess the possibilities for the introduction of DAB in various parts of the world. The meeting is expected to provide information about DAB system design, political and regulatory challenges, as well as the economic outlook for the new technology.

Advanced Technology Program Seeks Proposals —The Commerce Department announced today that it is seeking proposals for approximately $20 million in grants under its Advanced Technology Program (ATP) to help U.S. industry develop promising generic technologies. The announcement in the Federal Register set a deadline for receipt of proposals at 3:00 p.m. EDT, September 25, 1991. The ATP supports U.S. companies, either individually or in joint ventures, in developing precompetitive, generic technologies with significant commercial promise. The program is not restricted to any particular fields of technology. The program is administered by NIST. Further information and application packages may be obtained from the Advanced Technology Program, A430 Administration Bldg., National Institute of Standards and Technology, Gaithersburg, MD 20899.

Winners of TESLA Contest Announced — TESOFT has awarded prizes for the 1990-91 TESLA Models Contest. Prizes went to the individuals who contributed the best new models for the TESLA Block Diagram Simulator. Special recognition went to Tony Radice at General Instruments; Third Place to a TESLA user at Motorola who prefers anonymity; Second Place to Dr. Leo Montreuil at Scientific Atlanta and First Place to Dr. Joerg Friedrich at ANT Nachrichtentechnik in Germany. The new models are available free of charge on TESOFT's Bulletin Board System to all TESLA users.

INFO/CARD 21

NEW from KALMUS...



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- ★ 1-1000 MHz Broadband
- ★ 40 dB Gain
- ★ 10 dB Gain Adjust
- ★ Only 16 Pounds
- ★ MOS-FET Efficiency
- ★ 19'' Rack Adapters Included

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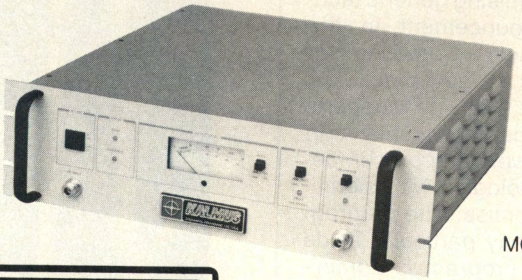
MODEL	POWER OUT	FREQUENCY RANGE	GAIN	SIZE (CM)	WEIGHT	AC LINE	U.S. PRICE $
700LC	1.5W CW	.003-1000 MHz	33dB	25x28x13	3.3kg	100-240V	$ 1,695
704FC	4W CW	.5-1000 MHz	33dB	23x18x09	2.8kg	100-240V	$ 2,195
706FC	6W CW	.5-1000 MHz	36dB	25x28x13	3.3kg	100-240V	$ 3,195
410LC	10W CW	.006-400 MHz	43dB	30x35x13	4.5kg	100-240V	$ 4,600
710FC	10W CW	1-1000 MHz	40dB	30x35x13	7.3kg	100-240V	$ 6,695
727LC	10W CW	.006-1000 MHz	43dB	48x46x13	8.5kg	100-240V	$ 7,750
711FC	15W CW	400-1000 MHz	40dB	30x35x13	5.5kg	100-240V	$ 3,620
720FC	25W CW	400-1000 MHz	40dB	48x46x13	8.6kg	100-240V	$ 5,995
712FC	25W CW	200-1000 MHz	40dB	48x46x13	8.8kg	100-240V	$ 7,350
737LC	25W CW	.01-1000 MHz	45dB	48x46x13	10.5kg	100-240V	$ 9,995
747LC	50W CW	.01-1000 MHz	47dB	48x46x26	26.5kg	100-240V	$22,500
707FC	50W CW	450-1000 MHz	47dB	48x46x13	13.0kg	100-240V	$ 9,995
709FC	100W CW	500-1000 MHz	48dB	44x48x18	22.5kg	100-240V	$19,990
722FC	200W CW	500-1000 MHz	50dB	44x18x31	41.5kg	100-240V	$31,900

Note: Models 727LC, 737LC and 747LC consist of two bands with one common input and output connector, switched with coaxial transfer relay, manually, or by remote. Switching speed 5 milliseconds.

MODEL 704FC



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RF calendar

September

11-14 **Radio '91**
Moscone Convention Center, San Francisco, CA
Information: Radio 1991 Convention Registration, 1771 N Street, NW, Washington, DC 20036-2891. Tel: (800) 342 -2460. Fax: (202) 775-2146.

15-19 **1991 Electronics Packaging Conference**
Sheraton Harbor Island Hotel, San Diego, CA
Information: International Electronics Packaging Society, 114 N. Hale Street, Wheaton, IL 60187-5113. Tel: (708) 260-1044. Fax: (708) 260-0867.

16-18 **International Conference on Analog to Digital and Digital to Analog Conversion**
Swansea, United Kingdom
Information: Secretariat, Conference Services, IEE, Savoy Place, London WC2R OBL, United Kingdom. Tel: 071 240 1871 ext. 222.

24-26 **13th Annual Electrical Overstress/Electrostatic Discharge Symposium**
Riviera Hotel, Las Vegas, NV
Information: 1991 EOS/ESD Symposium, c/o EOS/ESD Association, Inc., PO Box 913, Rome, NY 13440. Tel: (315) 339-6726. Fax: (315) 339-6793.

October

2-4 **The 13th Piezoelectric Devices Conference and Exhibition**
Kansas City Westin Crown Center
Information: Peter J. Walsh, Staff Vice President, Components Group, Electronic Industries Association, 2001 Pennsylvania Avenue, N.W., Washington, DC 20006. Tel: (202) 457-4932.

7-11 **13th Annual Antenna Measurement Techniques Association Symposium**
Clarion Harvest Hotel, Boulder, CO
Information: Sharon Sandera, 1991 AMTA Symposium, C/O NIST, Mailstop 813.05, 325 Broadway, Boulder, CO 80303 -3328. Tel: (303) 497-3302.

17-18 **Symposium on High Density Integration in Communications and Computer Systems**
GTE Laboratories, Waltham, MA
Information: Harry F. Lockwood, GTE Laboratories Inc., 40 Sylvan Road, Waltham, MA 02254. Tel: (617) 466-2786. Fax: (617) 890-9320.

17-21 **6th International Audio, Video, Broadcasting and Telecommunications Show**
Milan, Italy
Information: General Secretary IBTS, 20149 Milano, Via Domenichino, 11 - C.P., 15117-20150 Milano, Italy. Tel: (02) 4815541. Fax: (02) 4980330.

29-31 **RF Expo East**
Stouffer Orlando Resort, Orlando, FL
Information: Kristin Hohn, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, (800) 525-9154. Fax: (303) 773 -9716.

Technical Cellular
September 16-19, 1991, Madison, WI
Cellular Radio
October 21-24, 1991, Madison, WI
Information: University of Wisconsin - Madison, Francis P. Drake. Tel: (608) 262-2061. Fax: (608) 263-3160.

Analog MOS Integrated Circuits
September 16-20, 1991, Los Angeles, CA
Airborne and Spaceborne Radars: Surveillance, Tracking, and Imaging
September 16-20, 1991, Los Angeles, CA
Kalman Filtering
October 7-11, 1991, Los Angeles, CA
Data, Speech, Image, and Video Compression: Principles, Applications and Standards
October 28-November 1, 1991, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (213) 825-3344. Fax: (213) 206-2815.

Supplier Certification
September 17-18, 1991, Secaucus, NJ
September 19-20, 1991, New Orleans, LA
Information: Clemson University, Ms. Kay Barnett. Tel: (803) 656-2200. Fax: (803) 656-3997.

Infrared/Visible Signature Suppression
October 22-25, 1991, Atlanta, GA
Radar Cross Section Reduction
October 29-November 1, 1991, Atlanta, GA
Principles of Modern Radar
November 4-8, 1991, Atlanta, GA
Information: Education Extension, Georgia Institute of Technology. Tel: (404) 894-2547.

Cellular Radio Telephone Systems
September 30-October 2, 1991, Orlando, FL
Broadband Communication Systems
September 16-20, 1991, Washington, DC
Modern Receiver Design
November 4-8, 1991, Vienna, Austria
Introduction to Modern Radar Technology
September 25-27, 1991, Washington, DC
Lightning Protection
October 10-11, 1991, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Electromagnetic Compatibility
September 15-20, 1991, Canterbury, United Kingdom
Microwave Measurements
September 22-27, 1991, Canterbury, United Kingdom
Information: The Institution of Electrical Engineers, Savoy Place, London WC2R 0BL, UK.

Digital Signal Processing Workshop
October 15-17, 1991, Campbell, CA
Design Seminar: Principles of EMC
November 5-7, 1991, Mariposa, CA
Workshop for EMI Certificate Testing
October 9-10, 1991, Mariposa, CA
Seminar on the High Intensity Electromagnetic Radiated Fields (HIRF)
October 23-24, 1991, Mariposa, CA
Advanced HIRF Seminar for Commercial Flight Applications
September 24-26, 1991, Mariposa, CA
Information: CKC Laboratories, Registrar. Tel: (209) 966-5240. Fax: (209) 742-6133.

The VXIbus Seminar 2 Day Overview
October 1-2, 1991, Woodridge, NJ
October 1-2, 1991, San Diego, CA
Information: Testech, Ltd. Tel: (708) 554-1222.

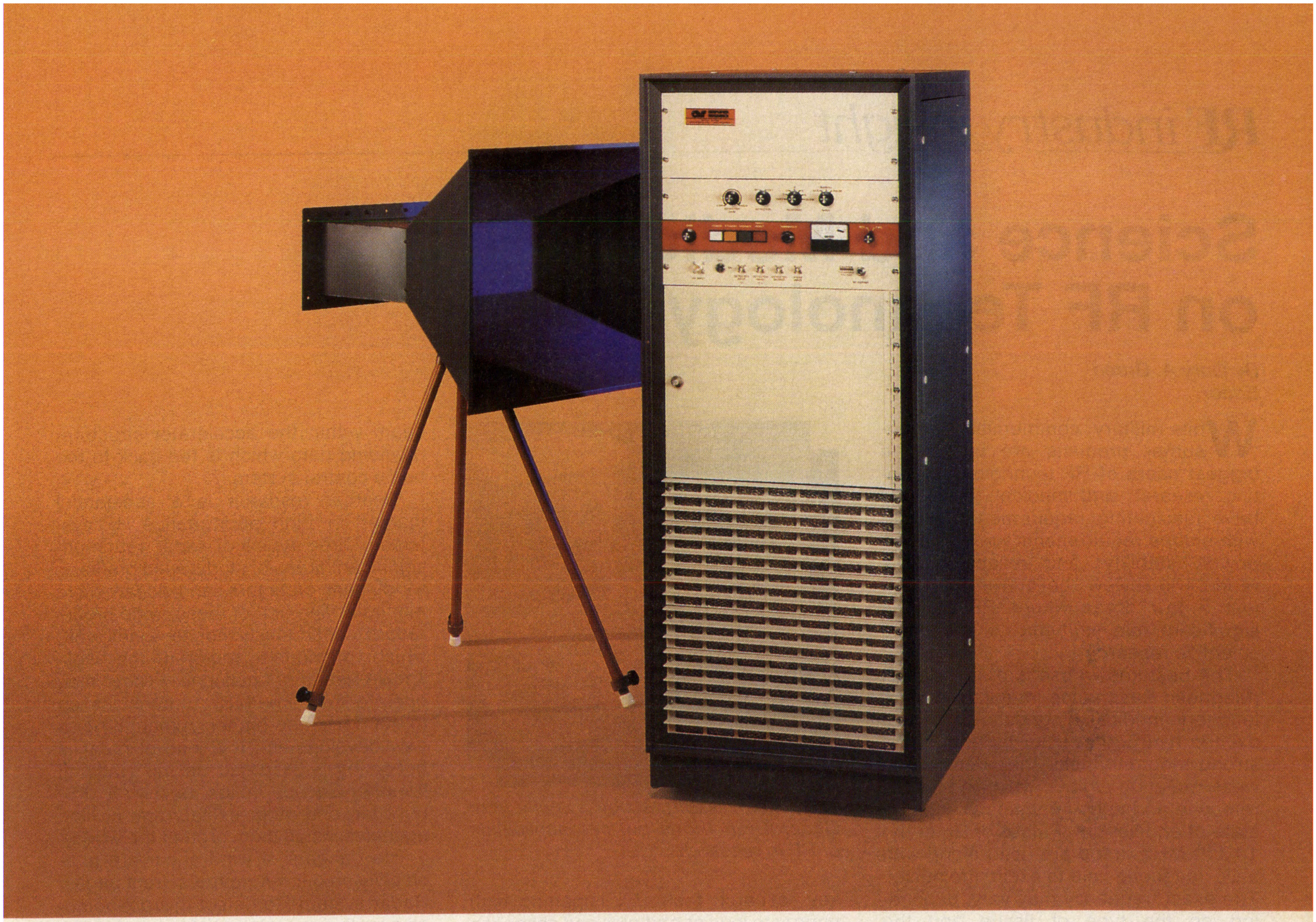
Digital Signal Processing Workshop
October 15-17, 1991, Campbell, CA
Information: Analog Devices, Maria Butler. Tel: (617) 461-3672.

Modern Digital Modulation Techniques
October 14-18, 1991, Spain
Broadband Telecommunications
October 14-18, 1991, Spain
Modern Digital Communications for Space, Satellite and Radio
October 14-18, 1991, Spain
Satellite Communication and Broadcasting
October 14-18, 1991, Spain
Combined Digital Coding and Modulation Techniques
October 14-18, 1991, Spain
Fast Algorithms for Adaptive Signal Processing
November 4-8, 1991, United Kingdom
Adaptive Signal Processing
November 4-8, 1991, United Kingdom
Array Signal Processing
November 4-8, 1991, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson, Box 910, S-612 01 Finspong, Sweden. Tel: 46 (0) 122-17570. Fax: 46 (0) 122-14347.

ELINT Interception
September 17-19, 1991, Syracuse, NY
ELINT/EW Applications of Digital Signal Processing
September 17-19, 1991, Syracuse, NY
Integrated EW
September 24-25, 1991, Syracuse, NY
ELINT/EW Data Bases
September 24-26, 1991, Syracuse, NY
Radar Vulnerability to Jamming
October 1-2, 1991, Syracuse, NY
Information: Research Associates of Syracuse. Tel: (315) 455-7157.

Basic Network Measurements Using the 8510B Network Analyzer
September 17-19, 1991, Atlanta, GA
September 30-October 2, 1991, Los Angeles, CA
Microwave Fundamentals
September 16-19, 1991, Los Angeles, CA
Designing for EMC
September 24-25, 1991, Washington, DC
Information: Hewlett-Packard Company. Tel: (714) 999-6700.

Practical EMI Fixes
September 16-20, 1991, San Diego, CA
EMC Design and Test
September 23-27, 1991, Las Vegas, NV
Information: Interference Control Technologies, Registrar. Tel: (703) 347-0030.



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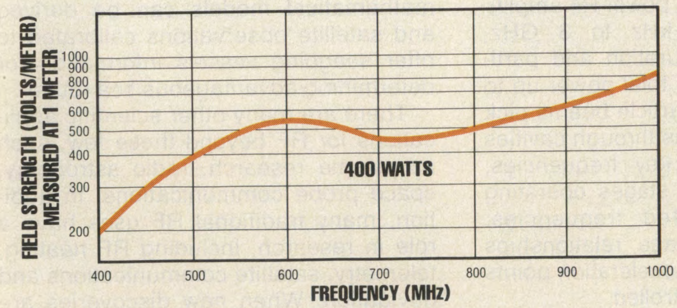

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Science and Medicine Rely on RF Technology

By Gary A. Breed
Editor

While military, commercial and consumer products are by far the biggest users of RF technology, scientific research and medical technology have fascinating requirements for RF, with unique requirements for sensitivity, power, stability, and reliability. This report will review the current status if just a few areas where RF plays a significant role in medicine and basic scientific research.

The best known use is probably MRI (Magnetic Resonance Imaging), a revolutionary medical diagnostic tool. MRI excites hydrogen nuclei by applying a strong magnetic field, then detects the weak radiation emitted when the field is removed and each nucleus returns to a state of equilibrium. Pulses of RF in the 10-110 MHz range are used along with a DC magnetic field to excite the nuclei. The exact frequency may vary, depending on the type of tissue and the sensitivity of the particular MRI system. Peak power can be up to 10 kW or more, or as low as 250 or 500 watts in some experimental low-power systems. MRI has been a substantial market for several power amplifier manufacturers.

This is a low duty cycle application, so RF power amplifiers must deliver high linearity pulses with substantial peak power, despite the low average power level. Nearly all systems use several frequencies, but most are narrowband, with excitation frequencies within a span of several kHz. In the detector or receiver, very weak decay emissions must be detected between the high power pulses. This creates another design constraint for sharp cutoff and low residual noise in the excitation signal, and fast switching and protection for low-noise detector circuitry.

Nuclear Magnetic Resonance (NMR) Spectroscopy, the original technology that led to MRI as a medical tool, is still widely used in biological, medical and organic chemistry research. Among these applications, General Electric reports that pharmaceutical companies and the petroleum industry are the most active areas at present.

In physics research, particle accelerators, cyclotrons and synchrotrons are key research tools. By imparting high energy to ions or particles and observing their interaction with other materials, scientists can study the fundamental principles of matter. At RF Expo East 1990, researchers at three different laboratories reported on their activities — Argonne National Laboratory, the Continuous Electron Beam Accelerator Facility (CEBAF), both in the U.S., and the KFA Julich research center in Germany. They identified three main areas of RF application.

This NMR Spectroscopy system from General Electric is but one application of RF in research.

First, acceleration of the ions or particles requires high power RF amplifiers, from 100s of kHz to 3 GHz, depending on configuration and particles accelerated, with total power up to tens of megawatts. Particle beams pick up energy as they pass through cavities excited with RF at key frequencies, typically with several stages operating at harmonically-related frequencies. Power level and phase relationships between all of these acceleration points must be precisely controlled.

Next, a variation of this method keeps the particle beam under control when injected into a storage ring. Maintaining and manipulating particle "bunches" requires frequency and phase control of the RF fields, as reported by Hovater and Abbott of CEBAF. Finally, monitoring techniques track the position of the beam within the accelerator structure, providing data which is fed back to the beam control system.

Weather research is a substantial market for high performance RF systems. Three types of radar represent significant areas of study: wind profilers, wind-shear detectors, and ocean wave monitors. The first of these, wind profile radars, continuously monitor winds aloft, which previously required the well-known weather balloons. They may operate from low MHz frequencies up to 400 MHz and higher. Although not a new development, this kind of radar is becoming widespread as the National Weather Service builds its central U.S. network, and other military and civilian agencies install them around the world.

The dangers of wind shear are periodically on our newspapers' front pages. Radar systems to detect this phenomenon are being installed at several major airports around the world, following initial testing in Colorado and Florida. Ocean wave measurement radars are another interesting application. Again, these have been in development for some years, but the needs of international commerce have required increased accuracy. Jon Adams of Jet Propulsion Laboratory reports that their NUSCAT system, a 14 GHz airborne radar measures wave motion and, indirectly, wind speed and direction. Using this data, mathematical models can be derived and satellite observations calibrated to offer seagoing vessels information for determining advantageous sea routes.

There are many other scientific applications for RF beyond these few, such as plasma research, radio astronomy, space probe communications. In addition, many traditional RF uses have a role in research, including RF heating, telemetry, satellite communications and navigation. When new discoveries are found in physics, chemistry, atmospheric sciences, biology and medicine, it is very likely that RF technology played an essential role. **RF**

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ETO understands the needs of OEMs, including responsive application engineering, fast prototyping, system integration assistance, and regulatory approvals. We combine rapid manufacturing ramp-up with strict quality standards, timely delivery, and international field service. We provide everything from small modules to complex systems with the "partnering" support OEMs require.

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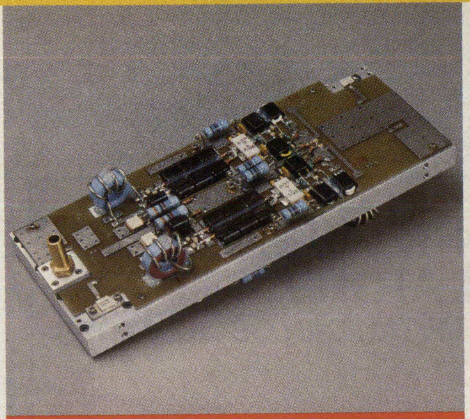
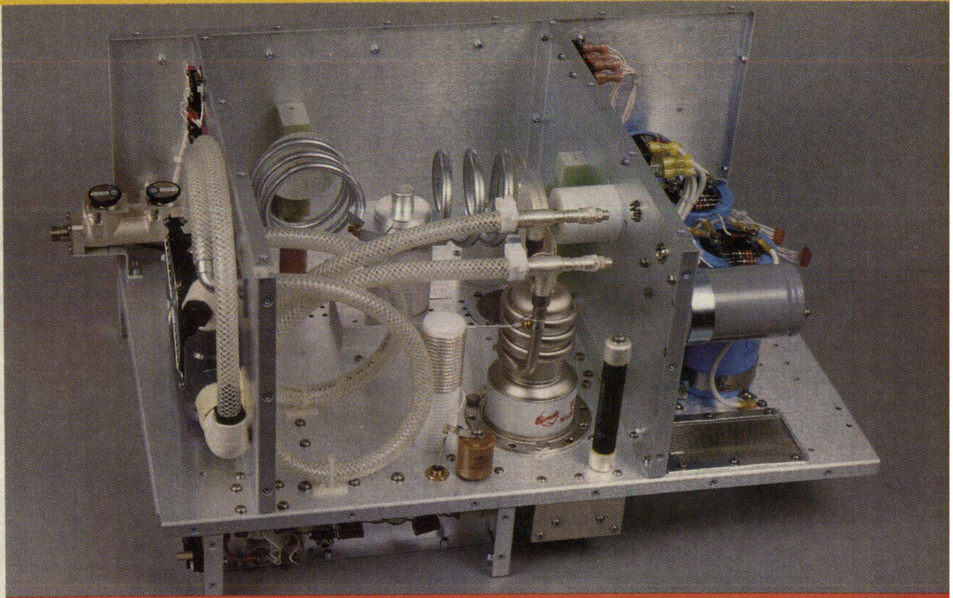
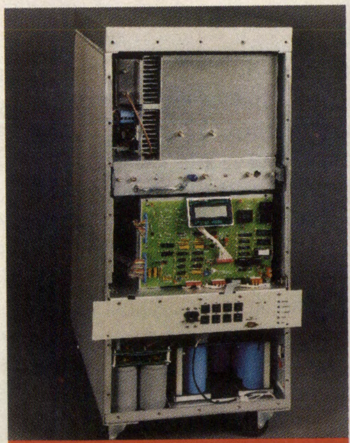


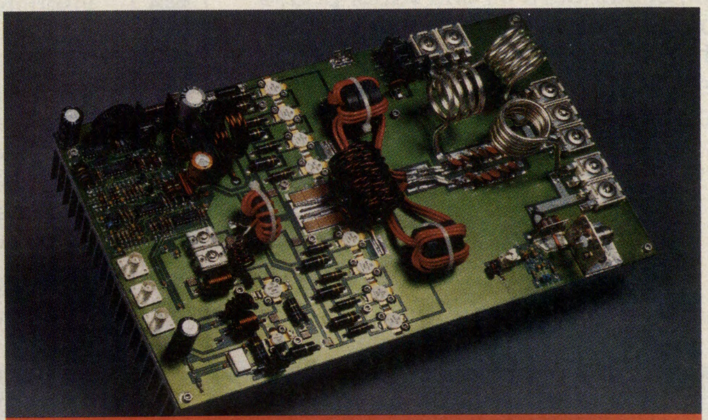
1 KW VHF POWER SUPPLY, ETO Model SGV-1D. *Water-cooled MOSFET design for plasma applications in the 50-150 MHz range. RF module only, or complete system including driver and power supply.*



5 KW RF POWER GENERATOR, ETO Model PG-5DW *for plasma and general industrial applications. RF deck of water-cooled version shown; available for any frequency to 90 MHz as a complete system including power supply and controller.*



20 KW RF LINEAR POWER AMPLIFIER, ETO Model PL-20DP. *Being delivered now to OEM's of high-field MRI systems. Available promptly for any frequency to 90 MHz. Microprocessor-based control system with integral LCD display permits local or remote management of the amplifier and its integral diagnostics.*



1 KW HIGH EFFICIENCY MOSFET HF POWER SUPPLY, ETO Model SG-1D *for plasma and other industrial applications. Available for 13.56, 27.12, or 40.68 MHz, or any frequency 2-41 MHz. RF module shown.*

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RF Power for Communications, Industry, Science & Medicine

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Very Wide Band Crystal Bandpass Filters

William B. Lurie, Consultant
Boynton Beach, Florida

The use of quartz crystals in bandpass filters is highly desirable because of their excellent stability and low losses. Unfortunately, the configuration of the electrical equivalent of such a crystal (a simple representation is shown in Figure 1) possesses two capacitors whose ratio, being never less than about 140, imposes a definite upper limit on the physically realizable bandwidth. In lattice filters without inductors in series or parallel with the crystals, that limit is the reciprocal of the capacitance ratio, or approximately 0.7 percent. By using added inductors, the maximum bandwidth becomes (theoretically) about 13 percent, but is actually somewhat less in practical filters.

Reference 1 describes the extreme steps needed to design and fabricate a sideband filter approximately 10 percent wide for an 88 kHz carrier, in the frequency range where available crystals have the lowest possible capacitance ratio. At higher frequencies, in the 4 MHz to 100 MHz range, where crystal filters are used extensively, somewhat higher capacitance ratios exist for the types of crystals used.

These limitations apply to lattice and half-lattice structures, whether image-parameter or modern insertion-loss designs, and even more so for ladder filters. It is not the purpose of this article to present a complete tutorial on crystal filters. Dr. Szentirmai's thorough treatise, Reference 2, explains all of the known methods of designing filters of all classes using crystals. These change somewhat as the bandwidth changes from "narrow" (less than about 0.8 percent), where added inductors are not needed, to "intermediate" and "wide' (up to perhaps 6 percent). As he says, "Finally, in very wide band filters, one can only realize a few selected attenuation peaks by embedding crystals in the branches of an LC ladder filter."

He also mentions transformations by

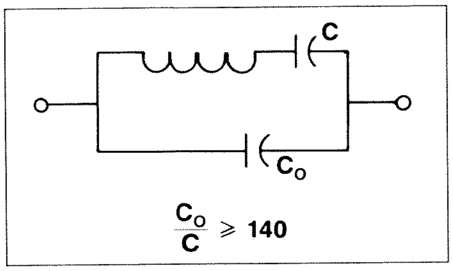

Figure 1. Typical equivalent circuit of crystal.

Poschenrieder (3), and Colin (4) whereby certain circuits can be modified to allow the substitution of a crystal for an equivalent LC structure. The most frequently seen of these, Figure 2 (see Reference 4), where a circuit realizable as a crystal has been incorporated into the portion of an LC ladder filter shown. Severe limitations are imposed on the lowpass in order to achieve a satisfactory capacitance ratio in the crystal, as indicated in the references. The two peaks of attenuation represented by branches L-C and L_1-C_1 must be quite close to each other and to the passband edge, in order for the capacitance ratio of the crystal (T_0/T_A) to be large enough to be achievable.

Such a lowpass filter can now be converted to a bandpass, using conventional transformations, and again severe limitations are imposed on the circuit elements and maximum bandwidth (see Figure 3 for central portion).

Another method of incorporating a crystal in a lowpass filter is available (Reference 4) which can also be con-

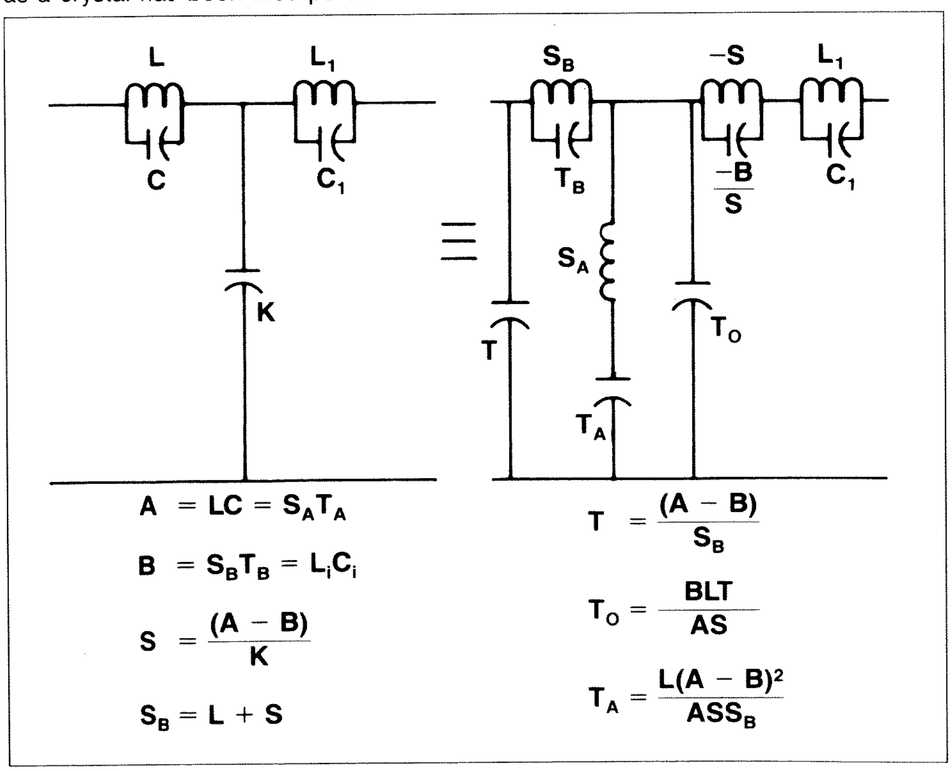

Figure 2. Most frequently used Colin transformation.

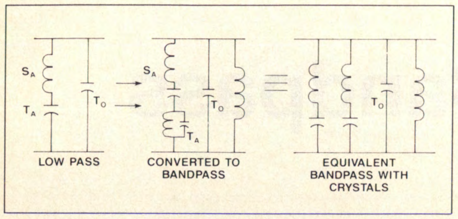

Figure 3. Conventional lowpass to bandpass transformation.

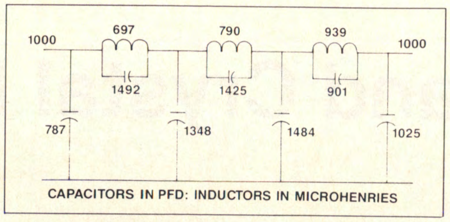

Figure 4. Lowpass filter before transformation.

verted to bandpass. An example follows. Figure 4 is a lowpass filter with cutoff at around 140 KHz, and Figure 5 shows an equivalent lowpass in two forms, one of which contains a crystal-like structure (5a) including a shunt inductor. These are, of course, exact equivelents. The filter of Figure 5b can, in turn, be converted to a bandpass centered at 138.3 KHz, again using a conventional transformation, and the central portion of the bandpass can be shown in two forms, Figures 6a and 6b, also equivalent. Examination of Figure 6a shows two circuits in the center, X and Y in dashed rectangles, each of which might contain a crystal, but the capacitance ratios are so low that the crystals would not be practical. Each of these circuits produces two frequencies of infinite attenuation, one below and one above the passband, explaining why the capacitance ratios are inadequate for crystals.

Referring to Figure 7a, the circuit of Figure 6b has been modified by simply interchanging the two parallel resonant circuits at the center. This leads to Figure 7b, where the central pairs of parallel-resonant LC circuits have been converted, each to an equivalent in which the familiar crystal structure reappears. This time, however, each circuit within the dashed lines makes two frequencies of infinite attenuation, both in the stopband on the same side of the passband. The result is that the capacitance ratio in each crystal is quite achievable in practice, near 300, even though the filter is 100 percent wide.

In the example given, if all inductor Qs are 300, the corners of the passband will be somewhat rounded, and the close-in peaks of attenuation will be somewhat reduced. These effects, due

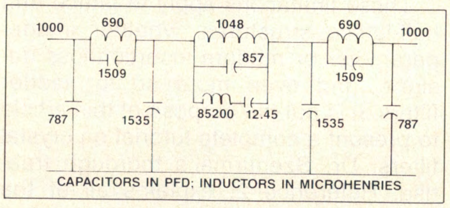

Figure 5a. Lowpass filter after transformation.

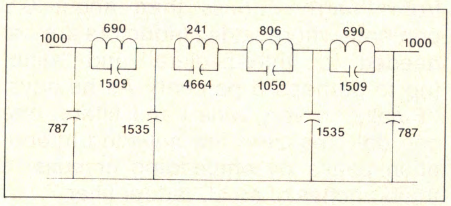

Figure 5b. Equivalent lowpass filter.

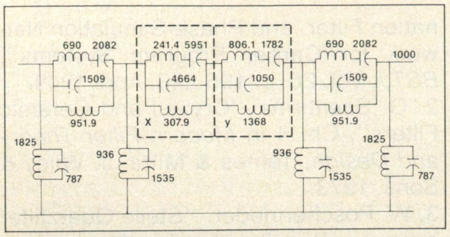

Figure 6a. Lowpass filter converted to bandpass.

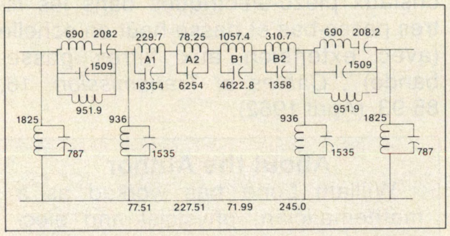

Figure 6b. Equivalent bandpass filter.

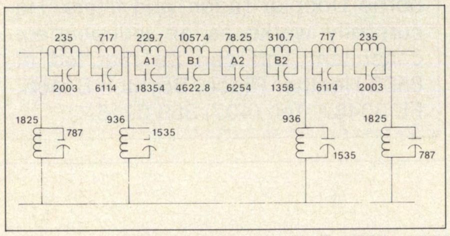

Figure 7a. Bandpass of Figure 6 with series circuits rearranged.

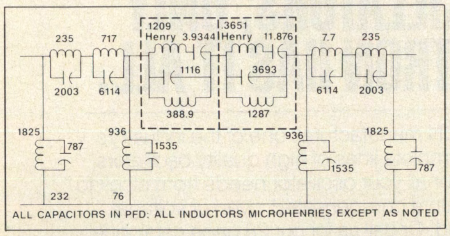

Figure 7b. Equivalent bandpass.

to component dissipation, are of course familiar in practical filters and can usually be tolerated. In Figure 8, the upper cut-off and stopband regions are shown. Curve A is the case where all inductor Qs are 300; Curve B has the two crystal Qs raised to 20,000. Near the cut-off, around 215 KHz, the curves are almost indistinguishable. In Curve B the closest-in peak has been sharpened very noticeably, and the farthest-out peak slightly. Unfortunately, some anomalous behavior in the stopband has appeared near the middle peak of the three, in Curve B, similar to that described by Szentirmai (2), for reasons not clearly understood at this time, although still under investigation. Although only the upper stopband is shown here, of course the lower stopband acts similarly.

One might conclude that, in this example, it is hardly worth the cost and effort to introduce crystals into such a filter, to achieve the relatively small improvements in notch depth near 227 KHz and 245 KHz (and in the lower stopband as well). That evaluation has some validity, but several points should not be overlooked. These attenuation peaks are relatively far from the passband, and so the filter without crystals still has a fairly sharp corner and well-defined stopband notches. For the same inductor Qs, these would deteriorate severely as the notches approach the passband edges. The curves shown already take into account inductor Qs

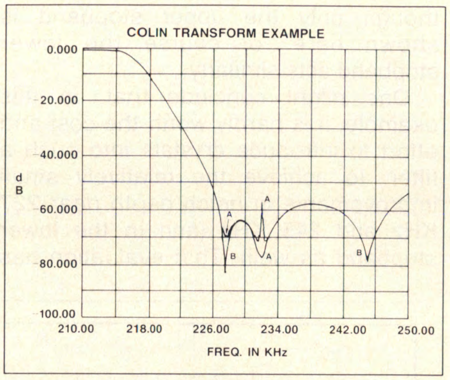

Figure 8. Amplitude response for filter of Figure 7b.

of 300, a figure which can not be bettered greatly for practical components. Finally, the method shown is entirely novel, in that it presents a methodology for introducing practical crystals into very wide filters for the first time.

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About the Author

William Lurie has worked as a mathematician, physicist and electronics engineer in a variety of fields, including magnetic compasses, X-ray tubes and measuring equipment, airborne Doppler radar, and filters. He currently works as an independent consultant and can be reached at 8503 Heather Place, Boynton Beach, FL 33437. Tel: (407) 369-3218.

Overtone Crystal Oscillators

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

Some overtone crystal circuit designs may be overly complicated for everyday standard applications. A reliable low cost circuit can be easily constructed using an RF MMIC amplifier, an overtone crystal and a minimum number of noncritical components. Two such basic designs will be presented.

The starting point is the Barkhausen requirement for oscillations. Using Figure 1, the open loop gain, G(s) × H(s) has to be larger than unity with a phase shift of 360n, where n is an integer (0, 1, 2, etc.). The simplest gain block, G(s), is an RF MMIC amplifier. These are available in inverting (basically 180 degree phase shift) and non-inverting (0 degree phase shift) configurations at low cost. These are essentially 50 ohm gain blocks (50 ohm input/50 ohm output) with typical bandwidths of over 1 GHz. The feedback path, H(s), is the crystal with its associated circuitry. The amplifier has to provide adequate gain to overcome the loss of the crystal circuit, and the crystal circuit has to compensate for any amplifier phase shift to obtain the required overall phase shift.

The two basic circuits presented here use either the inverting or noninverting amplifier configuration. The configuration of the available amplifier can be determined either from its S21 parameter or basic schematic given in the data sheet. The crystal used for the tests was a 91.25 MHz fifth overtone AT-cut type. Its approximate equivalent circuit is shown in Figure 2.

Inverting Amplifier Configuration

The Avantek MSA-0185 amplifier was used for the inverting amplifier configuration. From the data sheet it can be seen that it has a gain of about 19 dB and a phase shift of about 172 degrees at the frequency of interest. (An Avantek GPD-series hybrid amplifier could also be used eliminating two components: the bias resistor and the coupling capacitor). Thus, the crystal circuit has to be also of a phase inverting configuration to provide the required positive feedback.

Since it is desired to operate the crystal at, or very near, its series resonance, the phase shift through the crystal is essentially zero degrees. Thus, a phase shifting network has to be used to provide the needed approximate 180 degree phase shift. The technique described in Reference 1 was used. A full 180 degree shift cannot be obtained from either the Pi or Tee configuration, but it can be approached adequately close for this application. (The crystal will provide the small remaining required phase shift.) Four possible phase shaft configurations are possible depending on whether the 180 phase shift is approached from below (<180 degrees) or above (>180 degrees). For the Tee configuration, approaching the phase shift from below was selected. It uses only one inductor (a higher cost component) and two capacitors providing DC isolation (thus reducing the need for coupling capacitors). Since the phase shifter is also an impedance matching device, two configurations were tried: matched and with the circuit reversed (Figures 2a and 2b). The reversed configuration can be used since there is an excess of gain available. It has the advantage of unloading the amplifier and thus providing somewhat higher useful output (+6 dBm vs. +4 dBm at 16 mA total current).

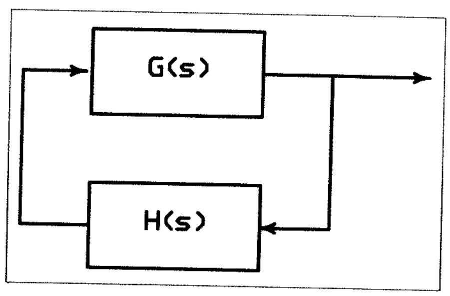

Figure 1. Feedback system.

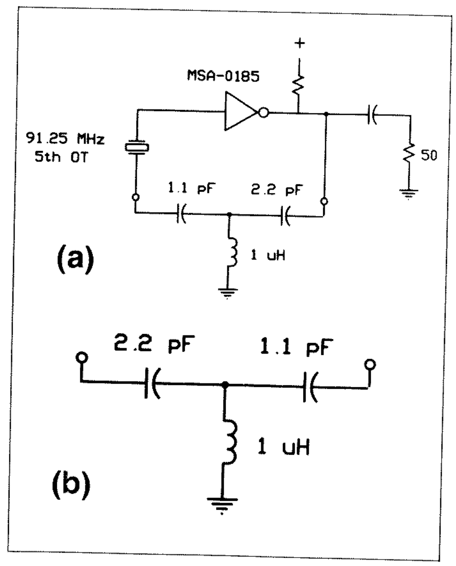

Figure 2. Inverting configuration.

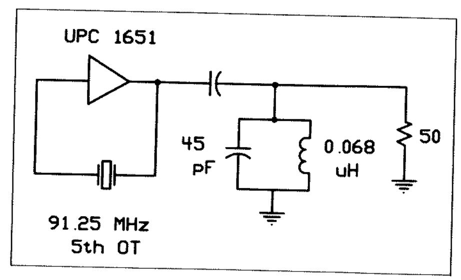

Figure 3. Noninverting configuration #1.

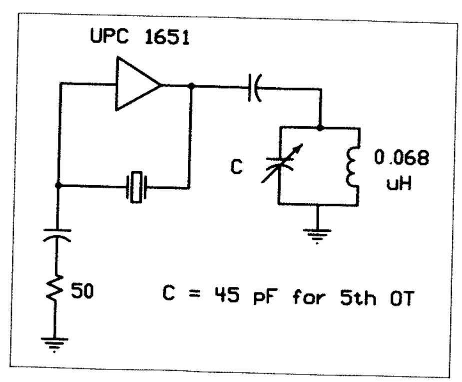

Figure 4. Noninverting configuration #2.

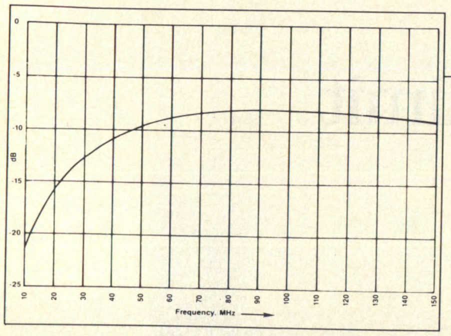

Figure 5. Tuned circuit, amplitude response.

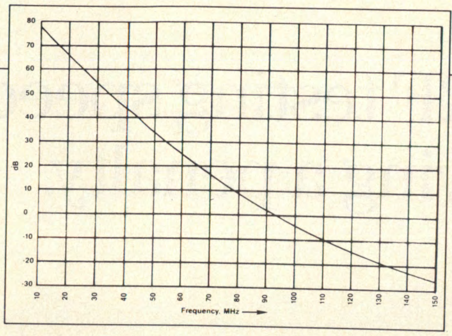

Figure 6. Tuned circuit, phase response.

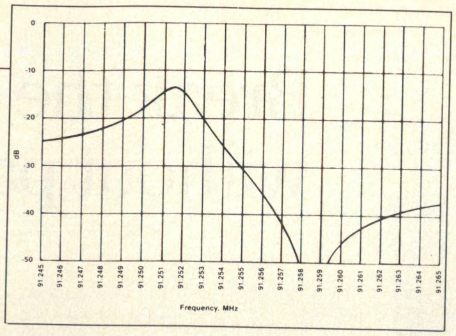

Figure 7. Total feedback, H(s), amplitude response at the fifth overtone frequency.

Noninverting Amplifier Configuration

Two configurations were tried using the noninverting NEC UPC1651 MMIC amplifier. Other noninverting amplifiers such as the Avantek INA-series can also be used. Since no phase inversion is required in the feedback circuit, the crystal can be used directly at series resonance. However, some means to prevent oscillation at the fundamental frequency and lower overtones is required. Everything else being equal, a circuit will tend to oscillate at the fundamental (or lower overtones) since the crystal series resistance is lower and the circuit gain is usually higher. Both of these effects increase the positive feedback at the lower frequencies. The two circuits are shown in Figures 3 and 4. In both configurations a parallel tuned circuit at the output of the amplifier, tuned to the desired overtone frequency, is used. This circuit reduces the open loop gain and introduces undesired phase shift at the unwanted frequencies (fundamental and lower overtones).

The circuit of Figure 3 was used to analyze the requirements. A similar analysis can be done for the circuit of Figure 4. The following approximation was made to simplify the calculations: the input and output of the amplifier is 50 ohms resistive. The effect of the tuned circuit on the feedback gain is shown in Figures 5 (amplitude) and 6 (phase). The response of the total feedback, H(s), at the fifth overtone frequency is given in Figure 7 (amplitude) and 8 (phase). It can be seen that near resonance the tuned circuit can be neglected and the crystal determines the circuit performance. At the fundamental frequency of the crystal (about 18.25 MHz), its resistance drops to about 20 ohms and the series capacity increases 25 times (both are approximations, but adequate for this analysis). The overall feedback response at the crystal fundamental frequency is shown in Figures 9 (amplitude) and 10 (phase). From Figures 5 and 6 we can see that

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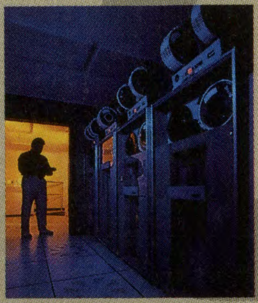

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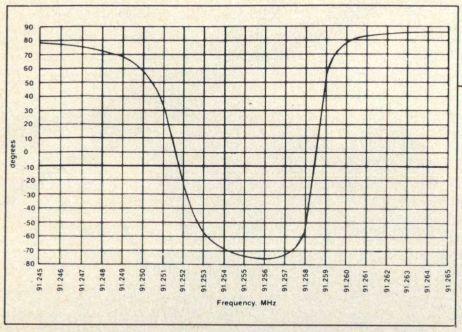

Figure 8. Total feedback, H(s), phase response at the fifth overtone frequency.

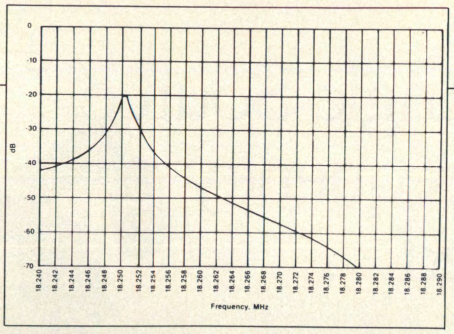

Figure 9. Total feedback, H(s), amplitude response at the fundamental frequency.

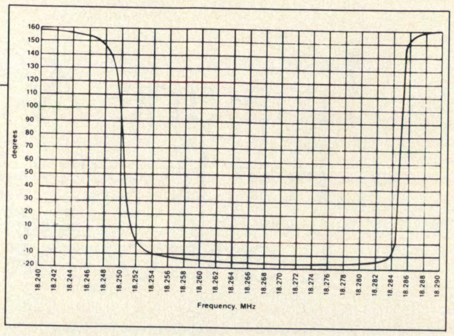

Figure 10. Total feedback, H(s), phase response at the fundamental frequency.

at the crystal fundamental frequency the response of the tuned circuit is down about 10 dB and the phase shift is about 68 degrees. To compensate for this phase shift (to provide 0 degrees total feedback phase shift necessary for oscillation) the crystal has to operate at a frequency higher than crystal series resonance. At this frequency the crystal amplitude response is down about 8 dB. Thus, the total feedback is down about 17 dB from the 5th overtone operating point, or about −27 dB, which is more than the gain of the amplifier, G(s), can overcome. Thus, no oscillation can take place at that frequency (the open loop gain is <1). The same analysis can be performed at the third overtone. The circuit will oscillate at the frequency at which the open loop gain is highest (and >1) at a 0 degrees phase shift frequency.

Both circuits (Figures 3 and 4) were built and had very similar performance. The circuit Q was about 36 percent of the unloaded crystal Q. The power output was −2 dBm. Circuit of Figure 4 had lower harmonic content and, possibly, lower phase noise. However, its crystal current and crystal power dissipation were much higher.

The circuits are quite uncritical. The parallel tuned circuit does not have to be very accurately tuned, as seen from Figures 5 and 6. If a very high resistance crystal is used, it can be compensated for by using a higher gain MMIC, such as one of the Avantek INA-series.

Using the circuit of Figure 4 and the same crystal, reliable operation at 3rd, 5th and 7th overtone could be obtained by varying capacitor C to tune the parallel tuned circuit to the desired overtone frequency. The output was constant at -2 dBm. Higher outputs can be obtained by using a higher power MMIC, but at the cost of increase of crystal power dissipation, unless a higher resistance crystal is used. Using a higher resistance crystal also decreases the reduction of the crystal Q by the circuit. However, the crystal Q may be lower to start with, unless its series inductance is increased.

The curves were drawn using a modified computer program described in Reference 2. A more thorough analysis can be performed by using the actual amplifier parameters and a more complex program. This is not considered necessary for the design of reliable circuit in this frequency range. For higher frequency crystal oscillators the techniques described in Reference 3 may be easier to implement. *RF*

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About the Author

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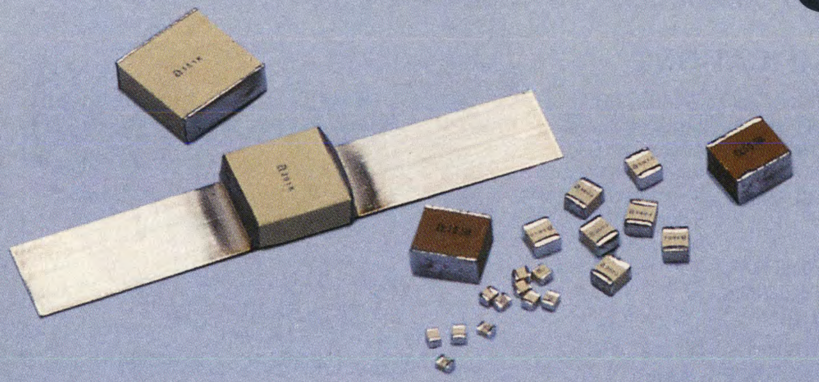

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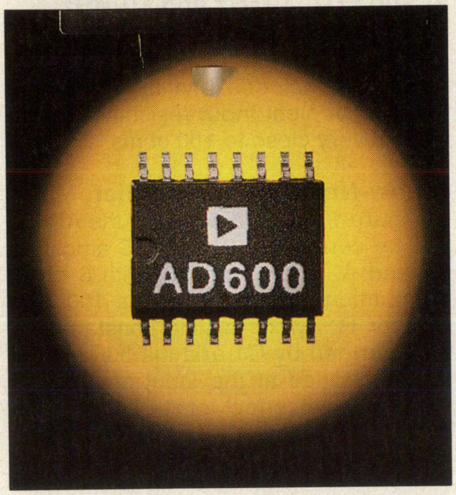
Low-Noise Amplifier with "Linear-dB" Gain-Control Simplifies AGC Systems

By Barrie Gilbert, Eberhard Brunner, and Bob Clarke
Analog Devices, Inc.

The design of wide-dynamic-range RF systems involves difficult tradeoffs between noise, bandwidth, distortion and power consumption. Voltage-controlled amplifiers (VCAs) are often used to reduce the dynamic range of a signal, but may not satisfactorily cope with these tradeoffs. A new VCA concept, called the "X-AMP", originally developed for use in medical ultrasound applications, combines very low noise (1.4 nV/√Hz) with a constant 3 dB bandwidth of 35 MHz, low distortion (-60 dBc to 10MHz) and low power consumption (125mW per channel). A feature of this technique is the provision of a fundamentally-accurate "linear-dB" gain-control law.



The key idea behind these amplifiers is the use of low-noise fixed-gain amplifiers, which can employ negative feedback to achieve low distortion and high gain accuracy, preceded by a broadband passive R-2R ladder network which provides an accurate 6.02 dB of attenuation between adjacent taps. A new technique allows these taps to be continuously interpolated, thus providing a smooth decibel-scaled gain function. Since the amplifier never has to cope with a large signal, distortion is further reduced, and because its gain is fixed the overall bandwidth is essentially constant. The resulting amplifier comes very close to the ideal VCA.

Two new ICs utilizing this principle have been introduced, both using a seven-stage ladder to provide a total attenuation range of 7 x 6.02 dB = 42.14 dB. In the model AD600, the gain of the amplifier is laser-trimmed to 41.07 dB. The gain control voltage V_G is applied to differential, high-impedance (approximately 15 Mohm) inputs, which have a scale factor of 32 dB per Volt. Thus, a change of 31.25 mV changes the gain by 1 dB. The gain for V_G=0 is at the midpoint, 20 dB; reduced to 0 dB when V_G=−625 mV and raised to 40 dB when V_G=+625 mV. Over this 0 to 40 dB range the gain is closely specified. When V_G exceeds these values, the minimum gain is −1.07 dB and the maximum is 41.07 dB. The peak output of these amplifiers is ±3 V for $R_L \geqslant 500$ ohm using the recommended ±5 V supplies.

The AD602 is very similar to the AD600, but each fixed-gain amplifier is trimmed to 31.07 dB, so each VCA section spans −10 dB to 30 dB (−11.07 dB to 31.07 dB max). The lower gain improves the output signal-to-noise ratio by 10 dB while the input noise remains 1.4 nV/√Hz.

The AD600 provides two identical, independent VCAs, which may be cascaded to provide 80 dB gain range. When their individual gain-control inputs are simply connected in parallel, the gain varies by 64 dB per Volt. However, in order to optimize the signal-to-noise performance over the entire gain range, a sequential arrangement which offsets the A1 and A2 gain control voltages can be used, in which the gain of A1 is first varied over its 40 dB range, while A2 operates at minimum gain. Thereafter, only the gain of A2 is varied to provide the second 40 dB segment of the overall gain range. We shall show how another modification to the gain-control function, using a small offset (3 dB), improves linearity.

The ladder network in both the AD600 and AD602 was chosen to have a value of R=62.5 ohms, in order to keep the thermal noise to within acceptable limits. The untrimmed input resistance to the

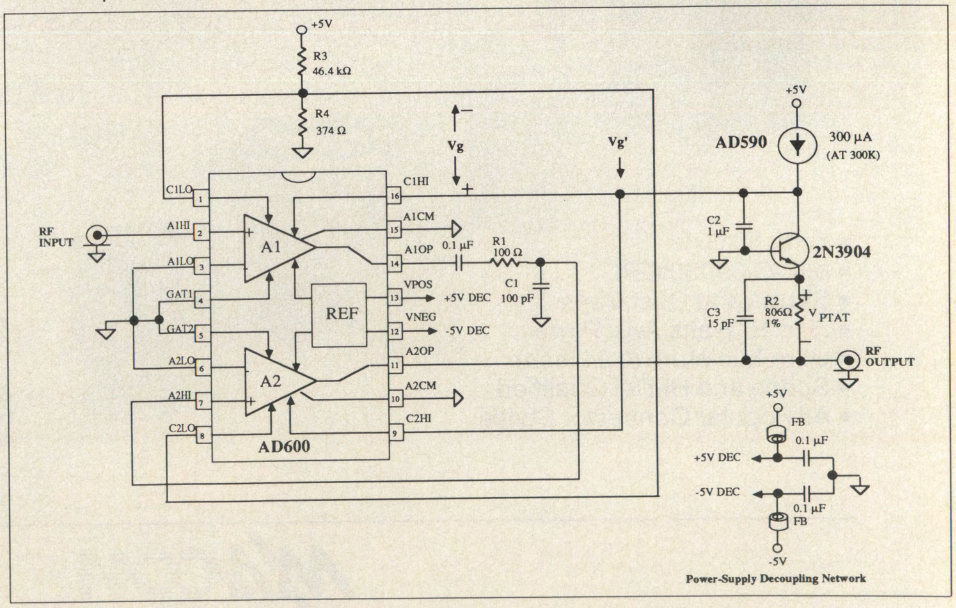


Figure 1. This accurate HF AGC circuit uses just three active components.

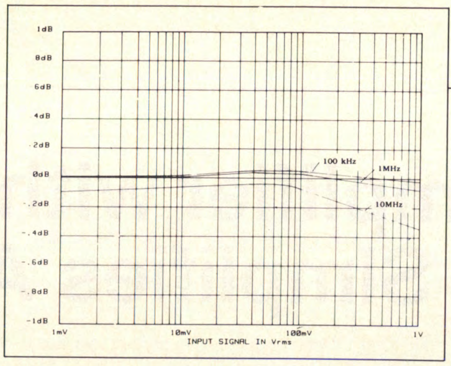


Figure 2. Output stabilization versus RMS input for sinewave inputs at 100 kHz, 1 MHz and 10 MHz.

ladder is thus 125 ohms. A padding resistor is included and laser trimmed for a convenient input resistance at the IC pins very close to 100 ohms.

A Low-Noise AGC Amplifier

Figure 1 shows how easily the AD600 can be connected as an AGC amplifier. A1 and A2 are cascaded, with 6 dB of attenuation introduced by the 100 ohm resistor R1, while a time constant of 5 ns is formed by C1 and the 50 ohms of net resistance at the input of A2. This has the dual effect of (a) moving the overall gain range to −6 dB to 74 dB, and (b) introducing a single-pole low-pass filter with −3 dB frequency of about 32 MHz. This is required to ensure stability at the maximum gain and reduces the bandwidth only slightly. The capacitor C2 blocks the small DC offset voltage at the output of A1 and introduces a high-pass corner at about 8 kHz.

A simple half-wave detector is used, based on Q1 and R2. The current into capacitor C2 is just the difference between the current provided by the AD590 (300 uA at 300K, 27 degrees C) and the collector current of Q1. In turn, the control voltage V_G is the time-integral of this error current. In order for V_G (and thus the gain) to remain stable, the rectified current in Q1 must, on average, exactly balance the current in the AD590. If the output of A2 is too small to do this, V_G will ramp up, causing the gain to increase until Q1 conducts sufficiently.

Since the average emitter current is 600 uA during each half-cycle of the squarewave a resistor of 833 ohms would add a PTAT voltage of 500 mV at 300 K, increasing by 1.66 mV/degree C. In practice, the optimum value of R2

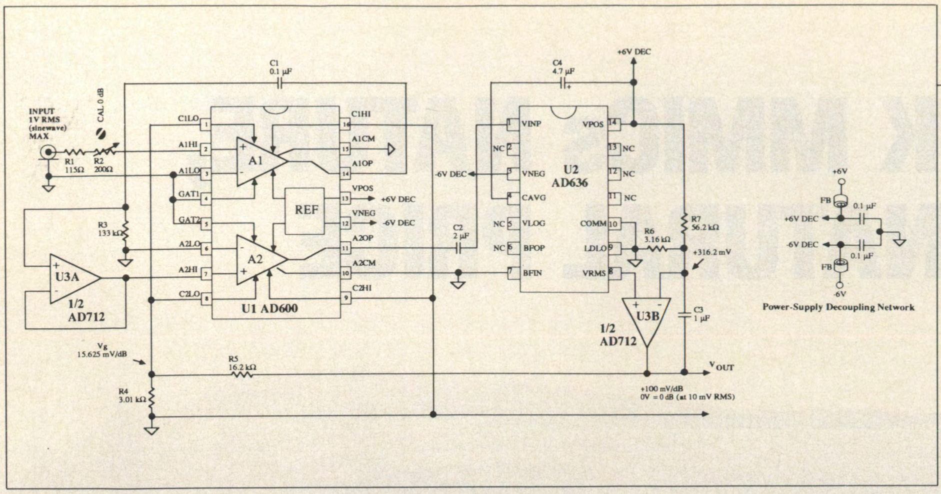


Figure 3. The output of this three-IC circuit is accurately proportional to the decibel value of the RMS input.

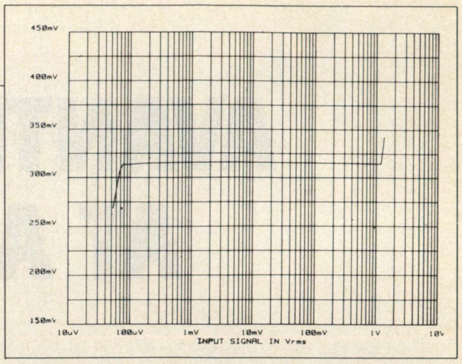


Figure 4. The RMS output of A2 is held very close tothe "set-point" of 316 mV for an input range of over 80 dB.

will depend on the type of transistor used, and, to a lesser extent, on the waveform for which the temperature stability is to be optimized. For the 2N3904 and sinewave signals, the recommended value is 806 ohms.

An offset of +375 mV is applied to the inverting gain-control inputs $C1_{LO}$ and $C2_{LO}$. Thus the nominal −625 mV to +625 mV range for V_G is translated upwards (at V_G') to −0.25 V for minimum gain to +1 V for maximum gain. This prevents Q1 from going into heavy saturation at low gains and leaves sufficient "headroom" of 4 V for the AD590 to operate correctly at high gains. In fact, the 6 dB interstage attenuator means that an input of 2 V_{RMS} would be required to produce a 1 V_{RMS} output at the minimum gain, which exceeds the 1 V_{RMS} maximum input specification of the AD600. The available gain range is therefore 0 to 74 dB (or, x1 to x5000). Since the gain scaling is 15.625 mV/dB (because of the cascaded stages) the minimum value of V_G' is actually increased by 6 x 15.625 mV, or about 94 mV, to −156 mV, so the risk of saturation in Q1 is reduced.

The emitter circuit of Q1 is somewhat inductive (due its finite f_T and base resistance). The addition of C3, determined experimentally to be 15 pF for the 2N3904, improves response flatness. Alternatively, a faster transistor can be used at Q1 to reduce HF peaking. The bandwidth is 40 MHz at 1.3 V_{RMS}. Figure 2 demonstrates the output stabilization for sinewave inputs of 1 mV to 1V_{RMS} at frequencies of 100 kHz, 1MHz and 10 MHz.

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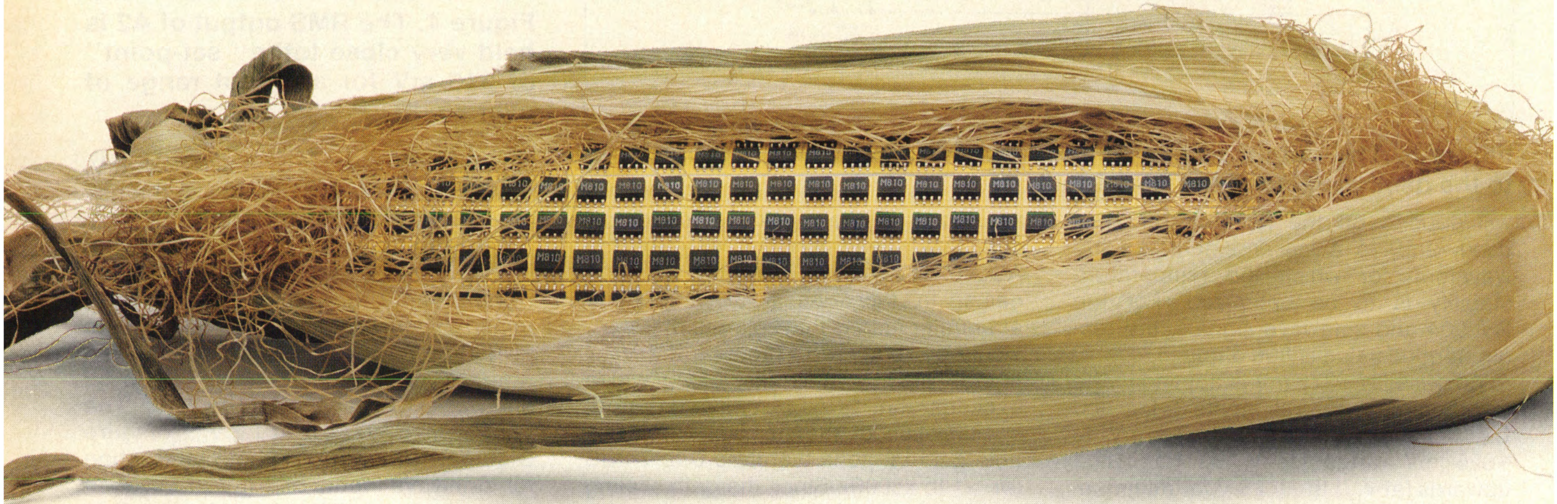

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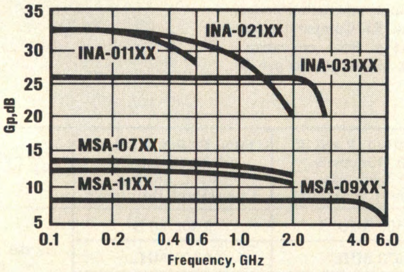
Power Gain vs. Frequency



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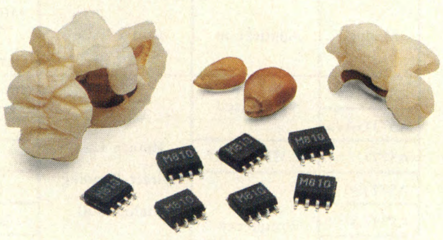

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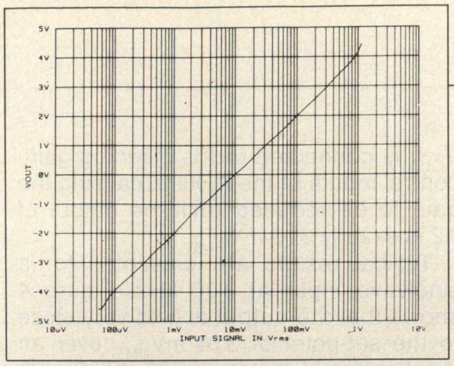

Figure 5. The decibel output of the circuit of Figure 3 is substantianlly linear over an 80 dB range.

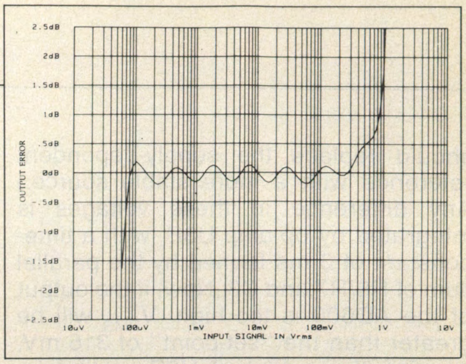

Figure 6. The data from Figure 7 presented as deviation from the ideal output given in Equation (1).

A Wide-Range RMS-Linear-dB Measurement System

Monolithic RMS-DC converters provide an inexpensive means to measure the RMS value of a signal of arbitrary waveform, and they also may provide a low-accuracy decibel-scaled output. However, they have certain shortcomings. The first of these is their restricted dynamic range, typically only 50 dB. More troublesome is that the bandwidth is roughly proportional to the signal level. For example, the AD636 provides a 3 dB bandwidth of 900 kHz for an input of 100 mV_{RMS}, but has a bandwidth of only 100 kHz for a 10 mV_{RMS} input. Its logarithmic output is unbuffered, uncalibrated and not stable over temperature; requiring considerable support circuitry.

All of these problems can be eliminated using an AD636 merely as the detector element in an AGC loop in which the difference between the RMS output of the amplifier and a fixed DC reference are nulled in a loop integrator. The dynamic range and the accuracy with which the signal can be determined are now entirely dependent on the amplifier used in the AGC system. Since the input to the RMS-DC converter is forced to a constant amplitude close to its maximum input capability, the bandwidth is no longer signal-dependent. If the amplifier has an exactly-exponential (linear-dB) gain-control law, its control voltage V_G is forced by the AGC loop to be have the general form:

$$V_{OUT} = V_{SCALE} \log_{10} \frac{V_{IN\,(RMS)}}{V_{REF}} \qquad (1)$$

Figure 3 shows a practical wide-dynamic-range RMS-responding measurement system using the AD600. It can handle inputs from 100 uV to 1 V_{RMS} with a constant measurement bandwidth of 20 Hz to 2 MHz, limited primarily by the AD636 RMS converter. Its logarithmic output is a loadable voltage, accurately-calibrated to 100 mV/dB, or 2V per decade, which simplifies the interpretation of the reading when using a DVM, and is arranged to be −4 V for an input of 100 uV_{RMS}, zero for 10 mV_{RMS}, and +4 V for a 1V_{RMS} input. In terms of Equation 1, V_{REF} is 10 mV and V_{SCALE} is 2 V.

As in the previous case, the two amplifiers of the AD600 are used in cascade. However, the 6 dB attenuator and low-pass filter found in Figure 1 are replaced by a unity-gain buffer amplifier U3A, whose modest bandwidth eliminates the risk of instability at the highest gains. The buffer also allows the use of a high-impedance coupling network (C1/R3) which introduces a high-pass corner at about 12 Hz. An input attenuator of 10 dB is provided by R1+R2 operating in conjunction with the AD600's input resistance of 100 ohms. The adjustment provides exact calibration of V_{REF} in critical applications, but R1 and R2 may simply be replaced by a fixed resistor of 215 ohms. This attenuator allows inputs as large as ±4 V_{PEAK} to be accepted.

The output of A2 is AC-coupled via another 12 Hz high-pass filter formed by C2 and the 6.7 kohms input resistance of the AD636. The averaging time-constant for the RMS-DC converter is determined by C4. The unbuffered output of the AD636 (at pin 8) is compared with a fixed voltage of +316 mV set by the positive supply voltage of +6 V and resistors R6 and R7. (Systems requiring greater calibration accuracy

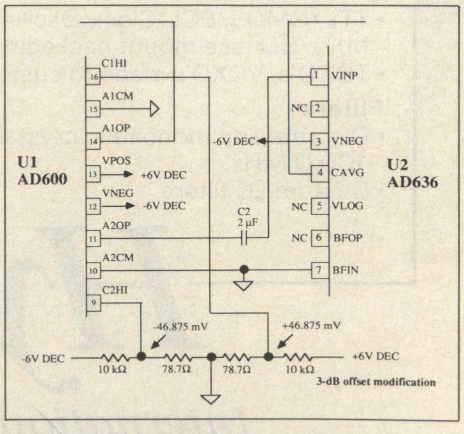

Figure 7. Simple method for reducing the gain-error ripple.

should replace the supply-dependent reference with a more stable source.) Any difference in these voltages is integrated by op-amp U3B, with a time-constant of 3 ms formed by the parallel sum of R6/R7 and C3. Now, if the output of the AD600 is too high, V_{RMS} will be greater than the "set-point" of 316 mV, causing the output of U3B — that is, V_{OUT} — to ramp up (note that the integrator is non-inverting). A fraction of V_{OUT} is connected to the inverting gain-control inputs of the AD600, causing the gain to be reduced until the output of A2 is exactly 316 mV_{RMS}.

Typical results are presented for a sinewave input at 100 kHz. Figure 4 shows that the output is held very close to the set-point of 316 mV_{RMS} over an input range in excess of 80 dB. Figure 5 shows the "decibel" output voltage. More revealing is Figure 6, which shows

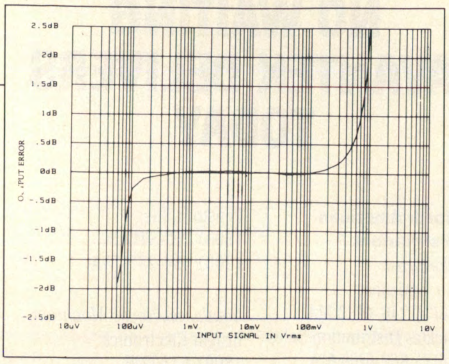

Figure 8. Using the 3 dB offset network, the rippleis reduced.

that the deviation from the ideal output predicted by Equation 1 over the input range 80 uV to 500 mV_{RMS} is within ±0.5 dB, and within ±1 dB for the 80 dB range from 80 uV to 800 mV. By suitable choice of the input attenuator R1+R2, this could be centered to cover other ranges from 25 uV-250 mV, to 1 mV-10 V, with appropriate correction to the value of V_{REF}. (Note that V_{SCALE} is not affected by the changes in the range). The gain ripple of ±0.2 dB seen in this curve is the result of the finite interpolation error of the X-AMP.

This ripple can be canceled whenever the X-AMP stages are cascaded by introducing a 3 dB offset between the two pairs of control voltages. A simple means to achieve this is shown in Figure 7: the voltages at $C1_{HI}$ and $C2_{HI}$ are offset by ±46.875 mV, or ±1.5 dB. Alternatively, either one of these pins can be individually offset by 3 dB and a 1.5 dB gain adjustment made at the input attenuator (R1+R2). The error curve shown in Figure 8 demonstrates that over the central portion of the range the output voltage can be maintained very close to the ideal value. The penalty for this modification is higher errors at the extremities of the range.

Three amplifier sections can be cascaded to extend the nominal conversion range to 120 dB, with the inclusion of simple LP filters of the type shown in Figure 1. Very low errors can then be maintained over a 100 dB range. ***RF***

For more information, please circle Info/Card #185.

About the Authors

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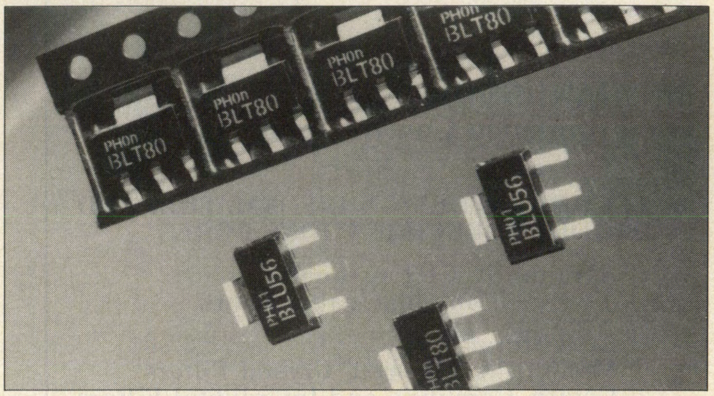
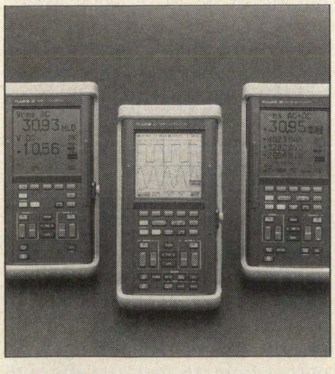
New Surface Mount RF Power Transistors

Philips Semiconductors has introduced the first RF power transistor in an industry-standard SOT-223 surface mount package. The BLT50, BLT80, BLU56, and BLU86 are all capable of dissipating 1 W of power and offer considerable pc board area and assembly time savings. The BLT50 and BLU56 are optimized for use at 470 MHz, and voltages of 7.5 V and 12.5 V, respectively. The BLT80 and BLU86 are designed for use in 900 MHz equipment, and voltage 7.5 V and 12.5 V, respectively. All the transistors feature high gain and efficiency in common-emitter class B amplifier stages and are particularly suitable as output or driver stage in cellular radio, mobile radio and other RF equipment. These RF transistors are available in bulk, or packaged on standard 12 mm bubble tape ready for use on SMD pick-and-place equipment. Pricing is in the $1.30 range for 1,000 piece quantities and delivery is 8-12 weeks ARO.
Philips Components
INFO/CARD #250



Digital Oscilloscope with Multimeter

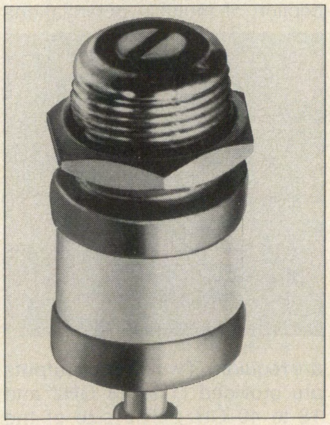
A new series of handheld service instruments combines a 50 MHz, 25 megasamples per second, dual channel digital storage oscilloscope with a feature-packed digital multimeter to bring extensive measurement capabilities in a rugged, sealed package to field service environments. The Fluke 93, 95 and 97 ScopeMeters from John Fluke Mfg. Co., Inc.



offer oscilloscope features with digital multimeter functions, while maintaining ease-of-use through such features as AUTOSET, 8 waveform and 10 set-up storage capabilities, 40 ns glitch capture time, combined display of meter results and waveforms, convenient menus and softkeys. The top model features a built-in signal generator, component tester, optically isolated RS-232 remote control and printer interface. The instruments have a U.S. list price ranging from $995 to $1595.
John Fluke Mfg. Co., Inc
INFO/CARD #249

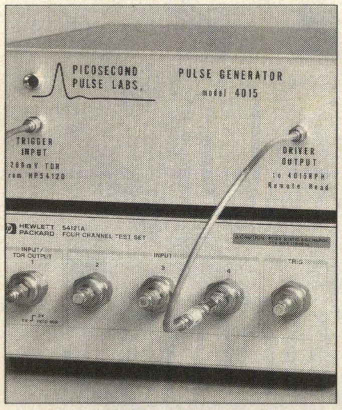
High Voltage Teflon Trimmer Capacitors

Voltronics is now producing a new line of sealed precision trimmer capacitors with Teflon replacing air as the dielectric. This increases the voltage ratings over 8 times to 1,000 DC working volts and 2,000 DC withstanding volts. The self-resonant frequency is over 2 GHz and the Q is over 3,000 at 100 MHz. They are internally O-ring sealed to withstand 40 PSI. The capacitance range is 1 to 9 pF tuning through 12 full turns with positive internal stops. TC is 0^{+50}_{-100} ppm/°C. The solid dielectric prevents ionization in the air gap, a major advantage in space, high altitude and high voltage RF applications. For quantities of 1,000, prices are $8.33 for the 9 pF ceramic case and $8.06 for the plastic case. Sample kits are available at half price.
Voltronics Corporation
INFO/CARD #248



Picosecond Pulse Generator

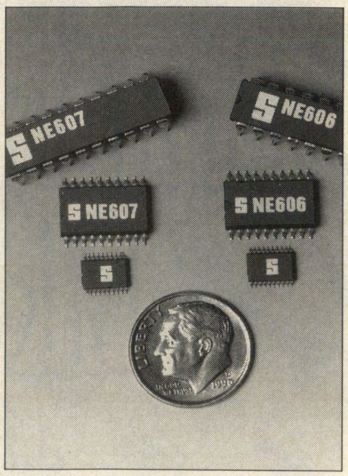
Picosecond Pulse Labs has announced the introduction of their Model 4015B Pulse Generator which produces -9 Volt pulses with an ultra-fast falltime of 15 picoseconds. With an accessory Model 5208 Impulse Forming Network attached, the 4015B can also produce extremely narrow, 22 ps wide impulses of -3.1 Volts.



The generator can be used as a companion TDR pulser for the Hewlett-Packard 54120 series of 20 to 50 GHz digital sampling oscilloscopes. When used as an amplifier, it will give a higher amplitude and faster pulse than available from the HP scopes alone. The fast 15 ps edge from the 4015B along with the built-in firmware "Normalization" capability of the HP54120 scopes will now provide 10 ps resolution for Time Domain Reflectometry and Transmission network measurements.
Picosecond Pulse Labs, Inc.
INFO/CARD #247

FM Receivers

The NE606 and NE607 are FM receivers which represent an advanced one-chip solution to FM Mixer-IF systems. These devices are designed to dramatically shrink the size and power consumption of portable communication systems such as cellular and cordless phones, wireless LANs and other communication receivers. One-chip integration helps reduce power consumption to 3.4 mA as well as lowering the voltage requirements to 3 V. Included on each chip is a mixer/oscillator, two operational amplifiers, IF amplifiers, limiter amp, voltage regulator and quadrature detector. The NE607 includes an automatic frequency control pin that enables users to stabilize IF



frequency for narrowband applications. Pricing in quantities of 100 is $3.57 for the NE606 and $3.68 for the NE607.
Signetics Company
INFO/CARD #246

Custom Grinding of Standard Cores

An unusual service for custom grinding standard ferrite cores to adapt them to meet specific requirements has been inaugurated by Ceramic Magnetics, Inc. This service, which can be performed on cores provided by the customer or Ceramic Magnetics, includes a variety of highly skilled operations. Among these are gapping, polishing, grading, modifying, and automatic sorting.

Ceramic Magnetics, Inc.
INFO/CARD #245

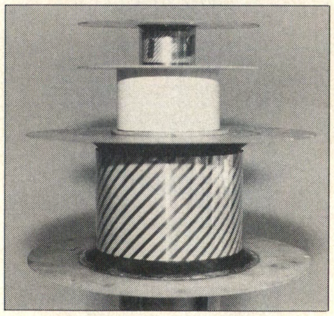
Monolithic Crystal Filters

OPT Industries, Inc. has announced the availability of two series of monolithic crystal filters. One series is centered at 10.7 MHz, and the other at 21.4 MHz. In each series, standard bandwidths are 7.5 kilohertz, 12 kilohertz, 15 kilohertz, and 30 kilohertz. The filters are available with two, four, six, and eight pole designs. Ultimate attenuation is 90 dB. The input and output impedance of the filters can be customized to meet specific requirements. Special features include low pass band ripple and excellent spurious response rejection. The operating temperature is −25 ° C to +75 ° C.

OPT Industries, Inc.
INFO/CARD #244

Broadband Slant-Linear Antenna

The latest Astron broadband antenna is the Model ODA-140 which combines only two antennas in a single compact radome to provide 0 dBi average omnidirectional gain over the wide bandwidth of 1 to 40 GHz. Antenna is ideal for applications where reception of multiple polarizations



are required. Two 50 ohm outputs are provided (1 to 18 GHz and 18 to 40 GHz). The size of the ODA-140 is 12 inches in diameter and 12 inches high.

Astron Corporation, Inc.
INFO/CARD #243

VXI Programmable Function Generator

Wavetek San Diego, Inc. has introduced a 12 MHz VXI function generator equipped with a sweep mode. Sweeping can be done in linear or logarithmic modes. Model 1370 can operate as a continuous, triggered, burst or gated signal source and can produce sine, triangle and square, square complement, DC and external width waveforms. Output levels are from 100 mV to 10 $V_{p\text{-}p}$ into a 50 ohm termination and 200 mV to 20 $V_{p\text{-}p}$ into an open circuit with 3.5 digits of resolution. Model 1370 is priced at $2,695.

Wavetek San Diego, Inc.
INFO/CARD #242

Field Strength Meter

The new Holaday HI-3012 is designed to measure RF electromagnetic fields to the new ANSI exposure limits. Two new probes measure to the lower limits of the ANSI uncontrolled environments for both electric and magnetic fields. The meter offers full-time automatic zeroing and a recorder output of 0 to 1 volts which corresponds to the meter reading. The HI-3012 comes with one E-field and one H-field probe, batteries, headset and instruction manual.

Holaday Industries, Inc.
INFO/CARD #241

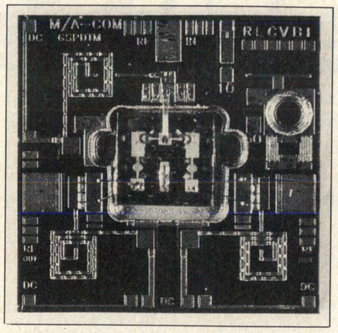
Power FETs

Excellent linearity and efficiency are the key features of the SU/S series of SST™ power FETs available from Microwave Technology. These devices have an output from 15 to 120 watts with 10 dB power gain when used in power amplifiers operating class A, B, or C in the 10 MHz to 500 MHz frequency range. Other features include operation at channel temperatures from −200° C to +200° C, no thermal runaway, high output VSWR tolerance, common source configuration.

Microwave Technology
INFO/CARD #240

PIN Diode Switch Chips

Using a multichip packaging technology, M/A-COM has devel-



oped a series of GMIC switches integrating a broadband bias network and a monolithic PIN diode structure. The GMIC SPDT switch chips are available as a decade bandwidth, 2-20 GHz switch (MA4MG201) and a lower insertion loss, 4-20 GHz switch (MA4MG202). Both switch chips have typical isolation values of 40 dB in the 8-14 GHz frequency band.
M/A-COM Inc.
INFO/CARD #239

Automated Microwave Tuning System

Applied Science and Technology, Inc. has introduced a new automated tuner. Designed primarily for microwave systems in production areas, the SMART TUNER™ allows the hands-free minimization of reflected microwave power. The device automatically positions three tuning stubs using a dedicated microprocessor controlled system in response to forward and reflected power signals received from the microwave power generator. The benefit to the user is a consistent power output to insure repeatability in process with minimum reflected power, therefore, maximum yield.
Applied Science and Technology, Inc.
INFO/CARD #238

Portable Spectrum Analyzer

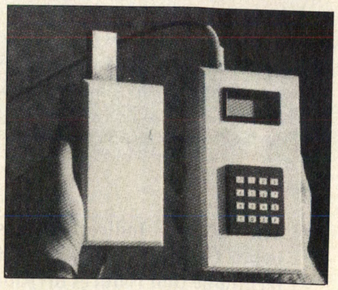
Hewlett-Packard Company announced it has extended the frequency coverage of its HP 8563A portable spectrum analyzer to 26.5 GHz from 22 GHz. An N-type input connector is standard for operation up to 26.5 GHz with no decrease in performance. This analyzer is one of the HP 8560 series portable spectrum analyzers which are built to the military-rugged specification of MIL-T-28800. The price starts at $32,000.
Hewlett-Packard Company
INFO/CARD #237

Synthesizer

The Combline Synthesizer can be used as a Built-In-Test or a Calibration Oscillator over the 500 MHz to 18 GHz frequency band. The Synthesizer can be tuned in .5 MHZ steps from 475 MHz to 525 MHz giving steps of 0.5 MHz x N where N is the combline multiple of 500 MHz. An option is available for adding a 4 way divider at the output. Another option adds a switch and an external RF input, thereby allowing the output to be switched between the external signal and the comblines.
EMF Systems, Inc.
INFO/CARD #236

Portable Reflectometer

Millimeter Wave Technology has announced a low cost hand-held portable Reflectometer for measuring the RF reflectivity of surfaces. The PR-12 standard



unit consists of a hand held RF measurement unit connected to a display unit via cable. The unit features microcomputer controlled electronics, a keypad interface and LCD display. The RF unit operates at 10.525 GHz and measures surface reflectivity over a 20 dB dynamic range with an accuracy of ±0.5 dB. The standard test aperture is linearly polarized. The unit is priced at $5,000.
Millimeter Wave Technology, Inc.
INFO/CARD #235

Monolithic Op Amps

Comlinear Corporation has announced its first ± 15 volt monolithic operational amplifiers. The CLC411 has a 320 MHz bandwidth and a 3000 V/μs sew rate and a superior gain flatness of ±0.1 dB to 30 MHz. The CLC430

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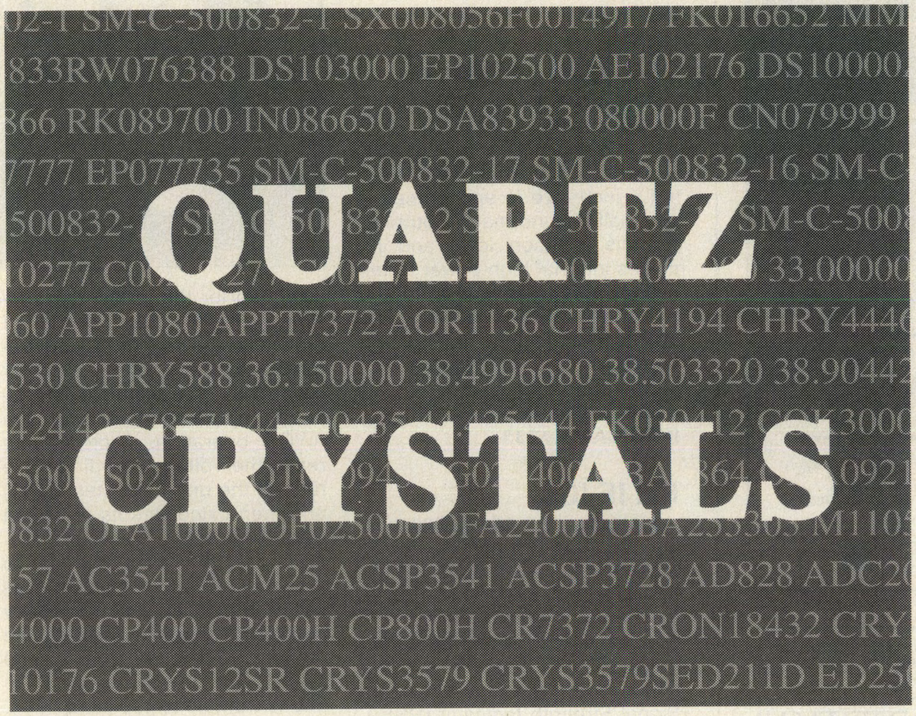

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has 55 MHz bandwidth and 2000 V/μs slew rate and is a current feedback opamp. The CLC430 offers 0.02 percent and 0.04° differential gain and phase for both NTSC and PAL frequencies.
Comlinear Corporation
INFO/CARD #234

250 Watt Isolator

This compact 50 dB isolator provides extra protection for 250 watt power amplifiers in the 270 MHz frequency range. The 3 × 4.5 × 1.5 inch size is approximately one half the normal "footprint" for this isolation level and includes internal high power terminations able to absorb up to 100 watts of reflected power. The insertion loss is under 0.5 dB, and the VSWR is 1.2 max. over a wide temperature range.
UTE Microwave, Inc.
INFO/CARD #233

FM IF ICs

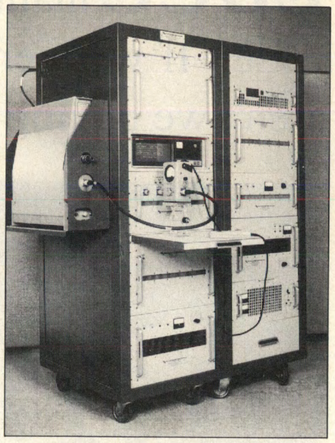
Motorola has added two new low-cost, low-power FM IFs to their list of RF communication ICs. The MC3371 an MC3372 have evolved from a family of RF devices which were introduced earlier. These devices have the added feature of a Received Signal Strength Indicator (RSSI). The MC3371 is designed for use with parallel LC quadrature detector components, while the MC3372 can be used with either a 455 kHz ceramic discriminator or parallel LC components.
Motorola Inc.
INFO/CARD #232

Low Loss Cable

The model 421-671, 0.141 inch diameter cable exhibits 30 percent lower loss than the MIL-C-17/1 limit at 1 GHz. Minimum centerline bend radius is 0.320 inches, and the cable can be used with standard RG 402/U connectors. The dielectric construction virtually eliminates the volumetric expansion normally encountered when heating solid PTFE dielectric cables, reducing VSWR failures in manufacturing and cracked joints in the field.
Storm Products Co.
INFO/CARD #231

Automated Wattmeter

Microwave Power Devices, Inc. has announced details of their automated wattmeter calibration system (AWCS). This system is



designed for the purpose of providing an RF output signal of precisely known power and frequency that is used in wattmeter calibration. The system's controller monitors and directs the AWCS operation. Microcircuitry reads the voltage standing-wave ratio of the unit under test (UUT) and then calculates its reflection coefficient at the test frequency. It also set the incident power to the UUT to a precise value. This value is set at any frequency between 1 and 400 MHz (at the output power levels from 1 W to 1 kW up to 30 MHz and 1 W to 500 W from 30 to 400 MHz).
Microwave Power Devices, Inc.
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Broadband Power Amplifier

The PA239 is an integrated RF amplifier featuring +16.5 dBm output power over the 10-2000 Mhz frequency range. Small signal gain measures 23 dB with gain flatness of ±0.5 dB typically from peak to peak. Noise figure is 5.5 dB with input/output VSWR's of 2.0:1 across the entire band. The amplifier is guaranteed to be unconditionally stable for any kind of source or load impedance terminations.
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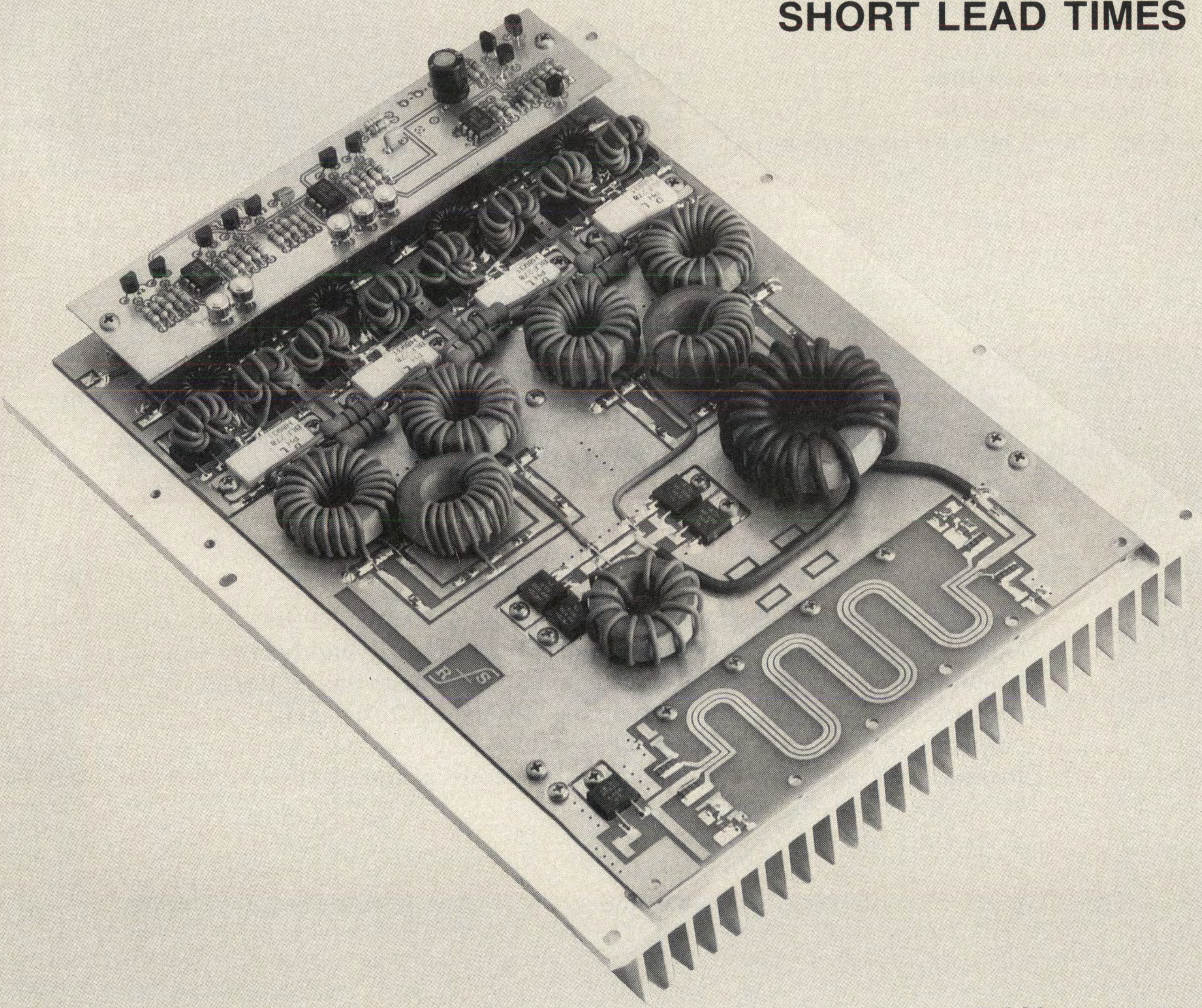

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Multifunction Counter

A Multifunction Counter with 5 Hz to 1.3 GHz bandwidth and TCXO timebase is now available from B+K Precision. The Model 1856 has frequency, period, period average, and totalize functions. The timebase has 0.5 PPM stability at 23 ± 5° C and 1 PPM stability from 0 to 50° C. For accurate low frequency measurements, the 1856 has period measurement capability. The period function's range extends from 0.285 μs to 0.2 seconds.
B+K Precision
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Calibrated Noise Sources

Noise/Com has introduced the NC346A, 6 dB ENR, and NC346C, 15.5 dB ENR. These devices are NIST traceable noise sources for noise figure measurement applications. The temperature and voltage stabilities are better than 0.009 dB/° C and 0.01 dB/% V, when operating from a +28 VDC supply due to the built-in regulator. On/off VSWR is less than 1.15:1 from 10 MHz to 5 GHz, 1.35:1 from 18 GHz to 26.5 GHz.
Noise/Com
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Special Products Section Piezoelectric Devices

Phase Locked Crystal Source

The new model PLX593 is a phase locked crystal source which features ultra low phase noise and exceptional signal purity. At a frequency of 812 MHz, phase noise at a 1 kHz offset is -120 dBc/Hz and at 10 kHz offset, -145 dBc/Hz. All reference subs, locking subs, and spurious are down -70 dB. Harmonics and sub harmonics are down 30 dB. The unit locks onto an input frequency of 10 MHz and offers an output power level of +10 dBm.
Techtrol Cyclonetics, Inc
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SAW Devices

Sawtek has introduced surface-mountable SAW filters, resonators, delay lines and other SAW devices with three different sizes of surface mount packages. These custom-developed packages are designed to satisfy high-volume, commercial surface mount assembly requirements while offering many features associated with military qualified SAW components. The surface mount package design features: a low height profile of less than 0.1 inch; one-shot, resistance weld hermeticity, high RF isolation between the input and output ports,; and proven, high-volume production compatibility.
Sawtek Inc.
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Ceramic Resonator

Model PL 3003 is a phase locked ceramic resonator oscillator providing two +15 dBm outputs at 1500 MHz. Isolation between outputs is 20 dB minimum, while harmonics are spurious at -35 dBc and -80 dBc, respectively. Phase noise, measured 100 kHz from the carrier, is -120 dBc. Operating voltage is -15 VDC at 200 mA maximum.
T and M Microwave, Inc.
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Digitally Compensated Oscillators

Models DC2200 AH and DC2210 AH are ASIC-based digitally compensated crystal oscillators. The two models use the ASIC for output frequency compensation and operate from a single, low power 5 Volt supply while maintaining a frequency stability of ± .2 ppm over the full operating temperature range. The frequency range is from 100 kHz to 25 MHz.
Murata Erie North America
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Temperature Compensated Oscillators

A new temperature compensated oscillator, Model X03003C, offers a standard ± 1 ppm stability

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over the temperature range of −35 to +75 degrees C. The standard frequency is 10 MHz with optional frequencies available between 8 and 15 MHz. The oscillator operates from a 12 volt supply voltage and its output options include sinewave, TTL and CMOS. Dimensions are 0.8 x 0.8 x 0.5 inches and the frequency adjustment is accommodated by an external potentiometer.
Piezo Technology, Inc.
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DIP TCXOs

Series CO-253 offers frequency stability of $\pm 2 \times 10^{-7}$ over 0 to +50 C or $\pm 1 \times 10^{-6}$ −55 to +85 C in a 24 pin DIP package. The aging rate is 1 x 10-6/year. Output is +7 dBm into 50 ohms with +13 dBm optional at any frequency in the 5 MHz to 140 MHz range. HCMOS/TTL compatible output is also available from 315 kHz to 50 MHz. Package is 0.8 x 1.4 x .39 inches.
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SONET Compatible Products

A broad selection of SONET compatible frequency control products is available from AT&T Microelectronics. These include 143-, 154-, and 157-Type Voltage Controlled Crystal Oscillators, Timing Recovery Units and SAW Filters. The products are available for the various SONET data rates and standards.
AT&T Microelectronics
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Crystal Oscillator Catalog

Catalog no. 10 from MF Electronics details their latest line of crystal clock oscillators. The catalog offers operating specifications, packaging options, environmental and mechanical information and ordering information. Copies of the catalog are also available on PC disk.
MF Electronics
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Near Field Coupling Between Signal Lines

By Vincent W. Greb
Ball Aerospace

This article has been written for the purpose of demonstrating the phenomenon of line to line coupling, or crosstalk between adjacent signal lines. It is also written to validate methods for predicting the magnitude of coupling for a given configuration. A line to line coupling experiment was performed on adjacent lines approximately one meter in length, in an attempt to support the theory with a macroscopic example. However, it will be shown that the principles established support EMC design practices and may be applied to other configurations, such as traces on a printed circuit board.

When signals on one line couple to another line, it is said that there is crosstalk between the lines. The term crosstalk is defined by MIL-STD-463A (1) as: "An undesired signal disturbance introduced in a transmission circuit by mutual electric or magnetic coupling with other transmission circuits."

Crosstalk is a significant issue and can result in many types of data errors unless design measures are taken to prevent it. With the evolution of new, faster integrated circuit technologies there also has evolved increased susceptibility to radio frequency (RF) noise. This is primarily due to two related factors: increased bandwidth of devices and smaller dimensions at the IC level. Increasing the bandwidth of a device allows it to operate faster but at the same time reduces its susceptibility threshold to higher frequency noise. As coupling increases proportionally with frequency, not only is more signal coupled onto traces due to the increased speed of operation, but the devices are now inherently more susceptible to such noise due to the smaller device dimensions. It will be shown that the voltages induced can be quite significant and thus, can be a contributing source to analog and digital signal errors. Among the more susceptible types of signals are 1) low-level analog and 2) single-ended digital signals.

Due to the fact that this experiment will be performed at frequencies no higher than 10 MHz and the lines will be very closely coupled, it is safe to assume near field conditions. The near field implies that the dominant coupling mode will be determined by the impedance of the noise source. A high impedance source will result in electric field, or capacitive coupling while a low impedance source will result in magnetic field, or inductive coupling.

The first portion of this article deals with a theoretical treatment of line to line coupling. The equations which will be used to predict coupling will be referenced, thereby defining the parameters which will govern the coupling mechanism. With this knowledge, a controlled experiment may be derived which will demonstrate the phenomenon. The second part of this article will be devoted to explaining the experimental setup and reporting the results. The experiment will be performed in an electromagnetically isolated environment to avoid errors which could confound the test results. It will give a demonstration of coupling between signal lines, the nature of which will be determined by parameters of the test setup. Finally, the experimental results will be discussed in terms of the theoretical predictions to prove the theory and to lay a basic foundation for predicting line to line coupling. Design techniques for reducing line to line coupling will also be discussed.

Theory

As stated, this experiment will only examine near field coupling between

Frequency (MHz)	Drive Volt. (Vrms)	Induced Volt. (mVrms)	Predicted Volt. (mVrms)
0.05	3.5	1.8	2.6
0.1	3.5	3.5	5.3
0.3	3.5	10.6	15.8
0.5	3.5	17.7	26.6
1.0	3.5	35.4	53.3
3.0	3.5	102.5	159.5
5.0	3.5	173.2	265.9
10.0	3.5	300.5	532.8

Table 1. Measured and calculated values for capacitive coupling.

Frequency (MHz)	Drive Curr. (mArms)	Induced Volt. (Vrms)	Predicted Volt. (Vrms)
0.05	224	0.054	0.056
0.1	224	0.079	0.113
0.3	224	0.157	0.338
0.5	224	0.243	0.564
1.0	224	0.443	1.1
3.0	224	1.4	3.4
5.0	224	2.3	5.6
10.0	224	4.4	11.3

Table 2. Measured and calculated values for inductive coupling.

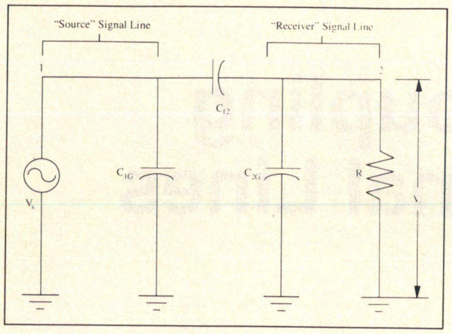

Figure 1. Electrical model for capacitive coupling (2).

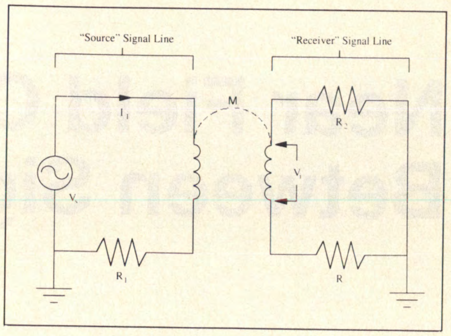

Figure 2. Electrical model for inductive coupling (2).

adjacent transmission lines. This analysis will encompass the majority of cases as electromagnetic coupling between lines usually takes place in the near field for several reasons. First signals driven on twisted pair transmission lines are limited to frequencies of approximately 10 MHz due to increased transmission losses at higher frequencies. This frequency corresponds to a wavelength of $\lambda > 10$ meters, resulting in a far field boundary ($\lambda/2\pi$) of 1.6 meters. In addition to this is the fact that signal cables are often routed in bundles which results in extremely close spacing ($r << 1.6$ m). Finally, the fact that the respective E-and H-field components of electric and magnetic fields drop off as a function of $1/r^3$, makes significant crosstalk very hard to achieve for signal pairs which are not closely coupled.

Electric field coupling is capacitive in nature; that is, it is a function of the capacitance which exists between two lines. This form of coupling is also termed "capacitive coupling." As the name "electric field" would imply, it originates from a high impedance source. The most extreme example of this would be a monopole antenna which is excited by an AC source. As the monopole is open at one end, it presents an extremely high impedance and is easily excited to an RF potential. The typical capacitance between a monopole and a ground plane is approximately 10 pF which corresponds to a capacitive reactance of > 1 Mohm at 14 kHz. It is therefore, very hard to drive significant amounts of current into the monopole. (This is only true until the monopole approaches a quarter wavelength.) Practically speaking, the magnetic field generated is negligible and the dominant field would result from a high impedance source which creates an electric field gradient emanating from the monopole. Consider the case where a number of twisted pair signal lines are being routed together, perhaps even within a single overbraid. Depending on the frequency of the transmitted signal, the pairs may be closely coupled for a considerable length. This will increase the capacitance between lines, which is directly proportional to equivalent surface area between adjacent conductors. Assume that the source is a 5 volt differential signal which is driving a relatively high equivalent impedance — on the order of thousands of ohms. This is done to minimize the power, and the resultant current is in the milliamp region. The potential for crosstalk definitely exists in this scenario and the equivalent model for capacitive coupling is shown in Figure 1. The induced common mode voltage (V_i) is defined by the following equation (2):

$$V_i = j\omega RC_{12}V_s \quad (1)$$

where ω = radian frequency
R = resistance of receiver to ground
C_{12} = line to line capacitance
V_x = voltage of source circuit

This equation shows very simply how the voltage induced on the receiver circuit may be varied. It does assume that the resistance "R" is much less than the capacitive reactance between lines plus the capacitive reactance between the receive line and ground. Were the converse true, the induced voltage would be defined by a voltage divider between C_{12} and C_{2G} as follows (2):

$$V_i = \left(\frac{C_{12}}{C_{12} + C_{2G}}\right)V_s \quad (2)$$

The experimentation to be performed however, will be designed based on Equation 1.

As the "R" and "V_s" values are often inflexible, the most practical way to minimize capacitive coupling between lines is to minimize either 1) the frequency of signal transmission and/or 2) the line to line capacitance. Often a minimum frequency for data transmission is also required, dictating that the lines be better isolated from one another. This is accomplished by running twisted shielded pair (TSP) transmission lines which, when properly terminated, greatly increase line to line isolation while at the same time, minimize the radiated electric field emissions from the signal pair.

Magnetic field, or inductive coupling, is the dual of capacitive coupling in that it is derived from a low impedance source and is a function of the mutual inductance which exists between two adjacent loops. The most extreme case of a low impedance source is the loop antenna driven by an AC source. The loop provides a very low impedance, especially at low frequencies. At higher frequencies, the impedance of the loop will be governed by its inductance; however, for low frequencies it is essentially a short circuit. Due to the small impedance, current will flow very easily but it will be very difficult to develop a potential across this load. Thus, the generated field will be primarily magnetic. Now consider the case where a relatively small load is being driven and a substantial current is required to drive such a load. The magnetic component of the generated field is much more significant than the electric field and a different type of coupling may be observed. An equivalent model for inductive coupling is shown in Figure 2 and the induced differential mode voltage (V_i) is defined by the following equation (2):

$$V_i = j\omega BA\cos\theta \qquad (3)$$

where ω = radian frequency
B = magnetic flux density
A = area of "receive" circuit
T = angle between adjacent loops

This equation is a more usable form of Faraday's Law of Induction which states that the induced electromotive force (EMF) in a circuit will be equal to the negative time rate of change of the magnetic flux which passes through this circuit. The negative sign, indicates that the induced EMF will be of such a polarity that the current generated from this source will set up a field to oppose the field which produced it. (While this fact will not be a concern in the coupling demonstration which will be performed, its existence is necessary to be consistent with the conservation of energy.)

The important parameters governing magnetic field coupling are obviously 1) frequency, 2) current, 3) areas defined by the respective circuit "loops" and 4) the orientation of the loops with respect to one another. Assuming that parame-

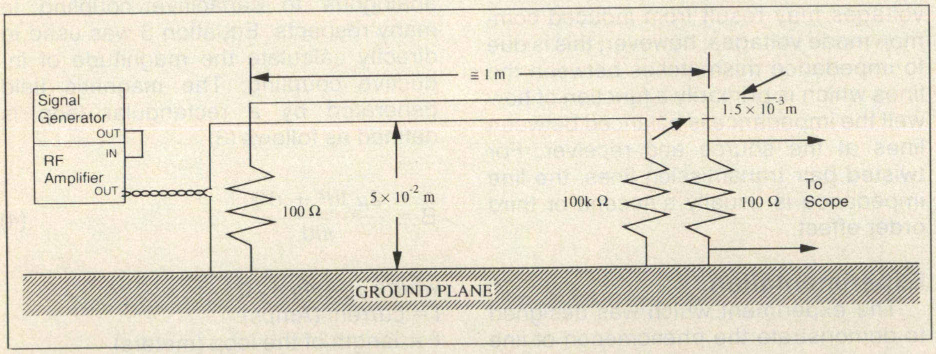

Figure 3a. Experimental setup for capacitive coupling.

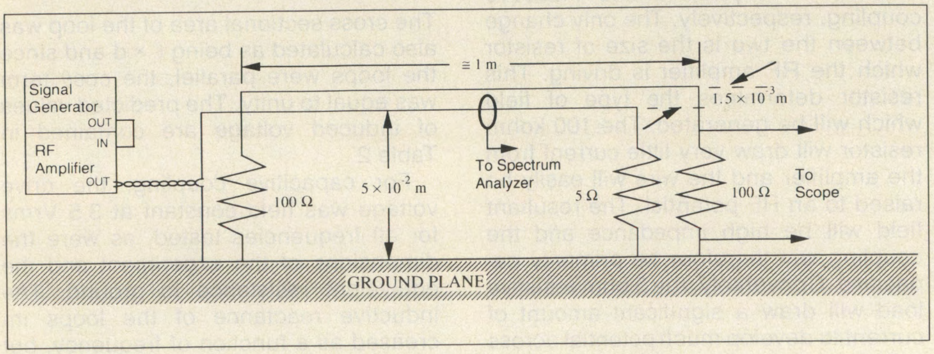

Figure 3b. Experimental setup for inductive coupling.

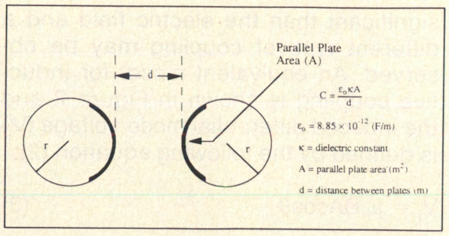

Figure 4. Equivalent capacitance for two parallel wires.

ters 1) and 2) cannot be altered significantly, consider once again the potential that exists for inductive coupling when a group of transmission lines are bundled together. They are essentially parallel, which maximizes the $\cos\theta$ term. However, there is a significant advantage to using twisted pair transmission lines. Twisting not only makes the characteristic impedance of the line more uniform but also decreases the loop area (i.e., the inductance) of the circuit this line will create. The result is a decrease in the potential magnetic emission and susceptibility level of this circuit.

It should be noted that twisting the wires will only reduce magnetic field coupling, not electric field coupling. This is due to the fact that magnetic field coupling results in a differential voltage induced in the circuit while electric field coupling results in common mode voltages induced in the circuit. Differential voltages may result from induced common mode voltages; however, this is due to impedance mismatches between the lines which is primarily a function of how well the impedance is balanced between lines at the source and receiver. For twisted pair transmission lines, the line impedance is usually a second or third order effect.

Experimentation

The experiment which was designed to demonstrate the phenomenon of line to line coupling is shown in Figure 3a and 3b for capacitive and inductive coupling, respectively. The only change between the two is the size of resistor which the RF amplifier is driving. This resistor determines the type of field which will be generated. The 100 kohm resistor will draw very little current from the amplifier and the wire will easily be raised to an RF potential. The resultant field will be high impedance and the coupling will therefore be primarily capacitive. Conversely, the 5 ohm resistive load will draw a significant amount of current to develop much potential across it. The resultant field will be relatively low impedance and the coupling will be primarily inductive.

As stated before, the amount of capacitive coupling will be a function of the capacitance which exists between lines. Line to line capacitance was determined by modeling the wires as a parallel plate capacitor and assuming the majority of the capacitance was due to the surface area defined by one third of the wire diameter times the length. The distance between the "plates" was simply two times the insulation thickness as the wires were taped together at approximately 10 cm intervals to ensure close and uniform coupling. The wire insulation had a dielectric of 5 and thus the line to line capacitance for a one meter length was calculated to be 68 pF. A picture of this model is shown in Figure 4. The actual capacitance was measured with a capacitance meter to be 48 pF and this value was used to calculate the induced voltage. Equivalent resistance to ground was chosen such that it would satisfy the requirements of Equation 1 and also would be at least comparable to the inductive reactance of the receive loop. Table 1 summarizes the results of the capacitive coupling in addition to the predicted value of the induced voltage.

A similar calculation was performed in order to determine the induced voltage due to inductive coupling, which is analogous to capacitive coupling in many respects. Equation 3 was used to directly calculate the magnitude of inductive coupling. The magnetic field generated by a rectangular loop is defined as follows (3):

$$B = \frac{2\mu_o I(\ell^2 + d^2)}{\pi \ell d} \qquad (4)$$

where $\mu_o = 4\pi \times 10^{-7}$
I = current (Amps)
ℓ = length of the loop (meters)
d = height of the loop (meters)

The cross sectional area of the loop was also calculated as being $\ell \times d$ and since the loops were parallel, the $\cos\theta$ term was equal to unity. The predicted values of induced voltage are contained in Table 2.

For capacitive coupling, the drive voltage was held constant at 3.5 Vrms for all frequencies tested, as were the dimensions of the experiment and the equivalent resistance to ground. The inductive reactance of the loops increased as a function of frequency, but as stated earlier, the resistors were chosen such that even at 10 MHz, they were comparable in impedance to the inductive reactance. Thus, the only parameter which was varied was the frequency. With all other parameters held (reasonably) constant, the induced voltage should increase proportionally with frequency, which is exactly what it did. Note that while the predicted values are consistently higher than the measured ones, they are all within 6 dB of the actual values, which is extremely good, and thus provide an excellent tool for predicting capacitive coupling.

The same is basically true for the inductive coupling experiment. Note that while the predicted and measured values are within 2 mV of one another at 50 kHz, the predicted value diverges to about 8 dB higher than the measured voltage at the higher frequencies. While still extremely close as far as predicting the magnitude of induced voltage, the disparity could be due to the fact that as frequency was increased, so did the load which the amplifier was driving due to increased inductive reactance. Thus, while the field was still fairly low impedance, the impedance of the source was increasing. Thus, the magnetic component of the field was decreasing while the electric field component was increasing. That notwithstanding, this prediction is also very close and is an excellent tool for predicting inductive coupling.

Design Recommendation and Conclusions

It is clear that the voltages induced onto the receive line are significant. Therefore, it is very easy to see how errors could occur, especially when dealing with low-level analog signals. Cable to cable coupling presents a significant problem due to the fact that cables can be very long and as a result, large loop areas and large amounts of line to line capacitance could be a problem unless techniques are implemented to minimize this during the design phase.

Inductive coupling can be reduced by simply twisting signal and return lines together, thereby minimizing loop area of the potential receive circuit. Routing cable bundles orthogonal to one another will minimize coupling between bundles. Similarly, at the board level, trace layers should be run perpendicular to adjacent trace layers. In addition, providing a return plane for each trace layer (as opposed to a single return plane for several layers) will greatly reduce both crosstalk and emissions.

Capacitive coupling can be controlled by minimizing the capacitance between adjacent lines. This may involve using a twisted *shielded* pair, providing that shields are well-terminated and exposed leads are minimized. Physically separating "noisy" from "sensitive" lines will also reduce potential crosstalk problems. Techniques discussed in the previous paragraph will also reduce capacitive coupling at the board level.

These are simply a few design guidelines which are derived directly from the equations which describe coupling. Many other methods exist to reduce coupling, however it is much easier if these techniques are implemented during the design phase. Investing a little more money up front is vastly superior to a costly EMC retrofit, which will inevitably create a substantial schedule impact as well.

Crosstalk is a very real and problematic phenomenon which is only becoming more significant as technology progresses. This experiment has demonstrated the phenomenon and the methods outlined herein for predicting coupling have been shown to be very accurate (within 8 dB). These methods can be extremely helpful in determining potential coupling magnitudes. understanding the mechanics of coupling will allow engineers to combat crosstalk and noise problems at the design level, which will ultimately save time and money and make systems inherently more reliable from an EMC standpoint. ***RF***

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3. D. Halliday, R. Resnick, *Physics Parts I and II*, John Wiley & Sons, 1966.

About the Author

Vincent Greb is an EMC Design Engineer for Ball Aerospace. His work involves EMI support for the Electroptics and Cryogenics Divisions. He may be reached at the Aerospace System Group, PO Box 1062, Boulder, CO 80306-1062. Tel: (303) 939-5940.

A Circuit Analysis Program for Filters

By Jack Porter
Cubic Corporation

This article describes the Third Place prize winner in the 1991 RF Design Awards Software Contest. One of many excellent filter programs entered in the contest, this one speeds analysis of active and passive filter networks.

General purpose circuit analysis programs such as SPICE are often used to evaluate filter performance, but they aren't the most convenient tools to use for that. These programs are large and often slow, the component models don't describe filter elements very well, the input data files are unwieldy, and the only useful output data obtained from the AC analysis are node voltages.

APFCAP is a special-purpose circuit analysis program which provides the necessary circuit elements and performs the the calculations necessary for frequency domain analysis of active and passive filters. It is a fairly small program, written in Fortran 77, which fits on a single floppy disk with ample room for data files. It will run on any IBM compatible computer and, although it executes quite rapidly, a math coprocessor is recommended.

A circuit can have a maximum of 50 nodes and 200 components, and output data can be tabulated at up to 300 frequencies. Nodes must be numbered consecutively (no number may be skipped), with node zero as ground. Each node must have a path to ground; op amp outputs and Norton amp inputs and outputs are such paths.

Figures 1a through 1h show the eight component types which are available. The components are connected to two to four nodes, depending on component type, and are characterized by an appropriate component value and, in some cases, an optional Q, gain-bandwidth or transfer capacitance. If this optional value is omitted, Q and G_{BW} are infinite and C_t is zero by default. Negative values are permissible for resistors, inductors and capacitors.

Figure 2 is a typical input data file. The file, titled ELFILA.DAT, and eight other examples are included on the program disk. The first line is a title, which may be up to 50 characters long. It can be blank, but must be the first line, not preceded by a blank line. The next

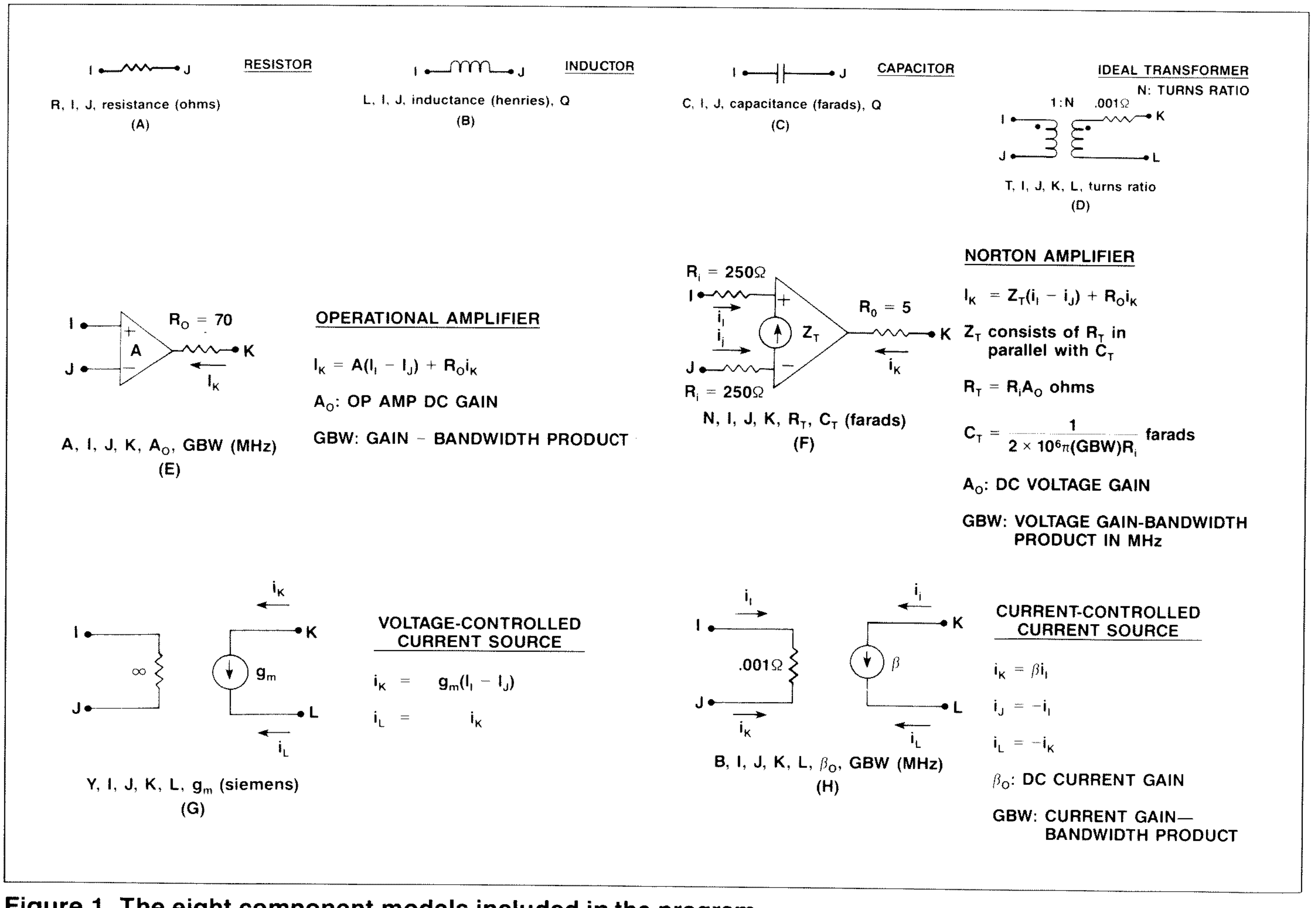


Figure 1. The eight component models included in the program.

```
PASSIVE ELLIPTIC FILTER — FB = 10.5 MHz
LIN, O, 10.5E6, 5E5/
R, 1, 2, 50/
C, 2, 0, 130E-12, 1000/
L, 2, 3, 0.73E-6, 300/
C, 2, 3, 100E-12, 1000/
C, 3, 0, 370E-12, 1000/
L, 3, 4, 0.85E-6, 300/
C, 3, 4, 120E-12, 1000/
C, 4, 0, 390E-12, 1000/
L, 4, 5, 0.94E-6, 300/
C, 4, 5, 27E-12, 1000/
C, 5, 0, 200E-12, 1000/
T, 5, 0, 6, 0, 2.0/
R, 6, 0, 200/
*, 1, 6/
Elliptic filter
N=7        Rho=3%        Theta=42 deg
FB=10.5 MHz     FH/FB=1.5     Amin=57 dB
Design max passband VSWR = 1.1
50 ohm filter terminations
2:1 transformer to 200 ohm load at output
```

Figure 2. Typical input file.

line contains frequency data. The first entry, the word LIN or LOG, specifies the type of frequency spacing. Next two are the lowest and highest freuqencies in Hz. The last entry controls the frequency interval. For linear spacing it is the frequency interval in Hz; for log spacing it is the number of frequencies per decade.

The following lines consist of a single letter denoting the component type, the nodes to which it is connected in terminal order I, J, K, L, as these terminals are defined in Figure 1, the component value, and the optional parameter is required. Figure 1 shows the data line format for each type of component.

The last line contains an asterisk followed by the input and output nodes. If the same node is designated for input and output, the impedance at that node and VSWR in a 50 ohm system are calculated; otherwise the voltage gain and phase shift at the output node are calculated with respect to the input. In the latter case, group delay is also calculated if linear frequency spacing is selected, as is usually done for the filter passband.

The input signal is a one amp current, thus the impedance at the input node is numerically equal to the node voltage. The voltage gain is a ratio of node voltages which are calculated from the

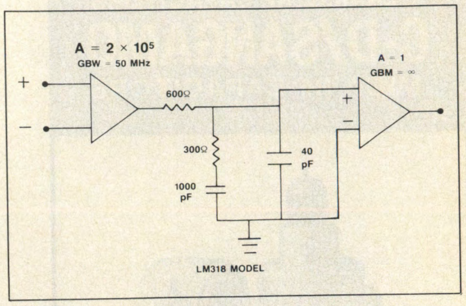


Figure 3. Modified op amp model to approximate an LM318.

network admittance matrix using the algorithm described in Reference 1. This is equivalent to applying a unit voltage source to the input node and calculating the output voltage.

Filters with finite transmission zeros, like elliptic and inverse Chebyshev filters, have phase discontinuities associated with them. Because of these discontinuities the numerical differentiation algorithm used for group delay calculation provides invalid results in the stopband of such filters. Fortunately, stopband group delay is rarely of any interest.

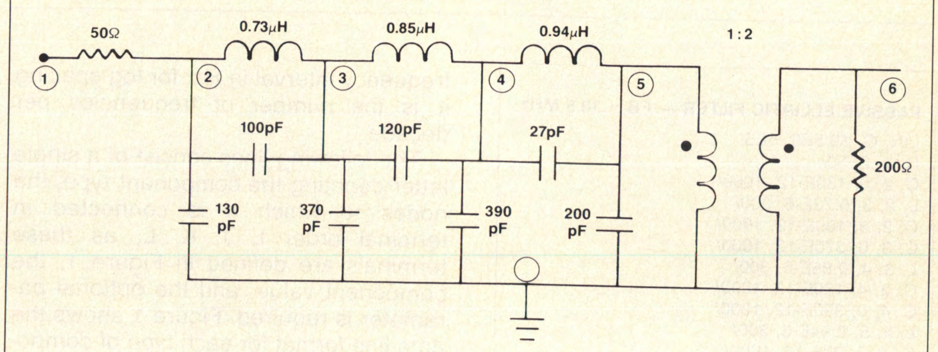


Figure 4. Elliptic filter example.

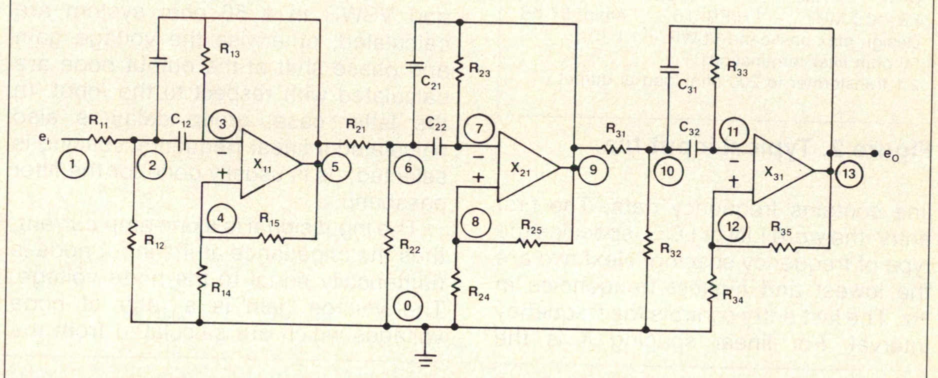


Figure 5. Three-resonator bandpass filter example.

Each line in the data file except the title must be terminated by a slash. Comments may be inserted in the file following the last data line or on any line following the slash.

Tabulated output data is displayed on the screen and, when the table is complete, may also be printed. Plots of gain in dB, phase shift and group delay or of VSWR are also displayed on the screen if the computer is equipped with an EGA, VGA or Hercules monochrome graphics card. Each pole remains on the screen until a key is struck.

Output data can be written to a disk file in a form that can be read by programs written in other languages, such as Basic and C, as well as by Fortran programs. The first line of this file contains information about the data table, including the number of frequencies and frequency spacing. Each following line consists of frequency, amplitude ratio, and phase shift data.

The circuit models used for resistors, inductors and capacitors are self-explanatory. Capacitor Q is defined as the reciprocal of the dissipation factor. The ideal transformer is sometimes used in passive filter models, and is essential in lattice filters. An actual transformer can be modeled using inductors and an ideal transformer. Reference 2 contains several examples.

The op amp parameters are DC gain and gain-bandwidth product in MHz. Open loop output resistance is 70 ohms. This model has a single pole in the open loop gain response. In cases where that is not sufficiently accurate, the actual open loop gain can be approximated by using this model and a network consisting of resistors and capacitors. Figure 3 shows such a circuit which closely approximates the open loop gain and phase characteristics of the LM318. File LM318.DAT plots its frequency response.

The 250 ohm input resistance used in the Norton amplifier model is approximately that of a diode with 200 microamps forward current. Positive or negative resistors can be inserted in series with the input terminals to adjust for other bias currents. The 5 ohm output resistance, which is approximately that of the LM359, can be changed in the same manner.

The transfer impedance, which is the input resistance multiplied by the voltage gain, is specified as a resistance R_t in parallel with a capacitance C_t. Approximate values for R_t and C_t can be calculated from the data sheet values for DC gain and gain-bandwidth product as shown in Figure 1g. If the input resistance is modified in the manner described above, the new value should be used in these calculations.

The other two available components are a voltage-controlled current source (VCCS) and a current controlled current (CCCS). Their main use is in high-frequency models for field-effect and bipolar junction transistors. The VCCS parameter is transconductance; the CCCS parameters are DC current gain and current gain-bandwidth product in MHz.

The examples include data files for three different filters. The first is a passive low-pass elliptic filter; Figure 4 is a schematic of that filter. Figure 2 shows the data file ELFILA.DAT used for calculating the passband response. The file ELFILB.DAT is used to calculate input impedance and VSWR and ELFILC.DAT for the stopband response. This is a 7-pole, 6-zero filter with 10.5 MHz 3 dB bandwidth and 1.5:1 shape factor. Nominal maximum passband VSWR is 1.1:1 and minimum stopband attenuation is 57 dB. It is designed for 50 ohm input and output impedance; the transformer at the output doubles the output voltage so that the voltage gain in dB is equal to the transducer power gain.

The second example, Figure 5, is a 3-resonator Chebyshev bandpass filter with a nominal 0.5 dB passband ripple, 250 kHz center frequency, 50 kHz bandwidth and 20 dB midband gain. The same filter was used as an example in Reference 3. The file CBPAF1.DAT shows the filter response with a 15 MHz G_{BW} single-pole op amp. In this case it is very close to the theoretical response. However, when this op amp is replaced by the op amp model of Figure 3 (CBPAF2.DAT) the passband is quite distorted. File CBPAF3.DAT shows the response of a redesigned filter which is much less sensitive to op amp characteristics.

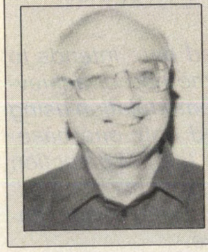
The file APFSET.PRT contains printer command codes. The code in the file is for an Epson LQ series printer, and it can be edited for compatibility with other printers. APFSET.PRT and LP77.EER, which generates run-time error messages, should be in the current directory.

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

References

1. S.D. Conte, C. de Boor, *Elementary Numerical Analysis*, Section 4.4, McGraw-Hill, 3rd ed., 1980.
2. S. Seshu, N. Balabian, *Linear Network Analysis*, Section 8.4, John Wiley & Sons, 1959.
3. J. Porter, "Stagger-Tuned Bandpass Active Filters," *RF Design*, March 1988, pp. 39-46.

About the Author



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A Single IF, Antilogarithmic, AM/FM Detector/IF Strip

By John C. Roberts
Northern Airborne Technology Ltd.

The circuit presented here intends to suggest an alternate method of achieving both AM and FM demodulation using a common IF strip and only one baseband op-amp IC. While the FM detection is straightforward, the recovery of amplitude modulation is somewhat unconventional, but very effective.

When designing communication equipment for overseas customers, one comes across an odd mixture of modulation schemes and bandsplits. Numerous orders and quotation requests have been presented to us asking for AM/FM non-military receivers and transceivers. One such application is in SAR (Search and Rescue) operations where both AM aircraft bands as well as FM marine frequencies have to be monitored. While planning the design of a SAR receiver, I did not wish to have two individual IF amplifiers, crystal filters and detectors in order to service both AM and FM modes. I felt that it would be very advantageous to come up with a scheme for simultaneous detection of both AM and FM modulation. This would allow several desirable features to be inherent in the design:

1) Low Cost: Unlike many of our systems, price was of concern as many offshore companies market functionally similar receivers.

2) A minimum of adjustments during alignment and calibration. This would minimize time spent in test and inspection.

3) Ease of layout: With the increased board space an unconditionally stable layout could be achieved.

4) No AGC requirement for the AM detector. (More on this later).

5) Low distortion recovery of both AM and FM, even simultaneously.

Circuit Description

The RF portion of the circuit is based on the SA604 family of FM IF modules. It has proven itself to be one of the most versatile FM IF/Detector ICs on the market today.

FM Detector

As far as the FM detector is concerned, the circuit is very conventional and its application is covered well in the Signetics application notes for this device. The layout is very conservative with as much use of groundplane as was possible. Detection is with a conventional FM quadrature detector. Sensitivity is typically less than 0.3 uV for 12 dB Sinad with the input matched to the 1.5 K input impedance of the mixer.

AM Detector

As the SA604 uses a limiting amplifier for the majority of its IF gain, another method had to be found to recover the amplitude information.

The SA604 has a current source RSSI pin which gives a DC current output that is proportional to the logarithm of the input signal. For instance, with a 180K resistor from this pin to ground, a voltage of 1 V/10 dB increase of signal strength can be observed.

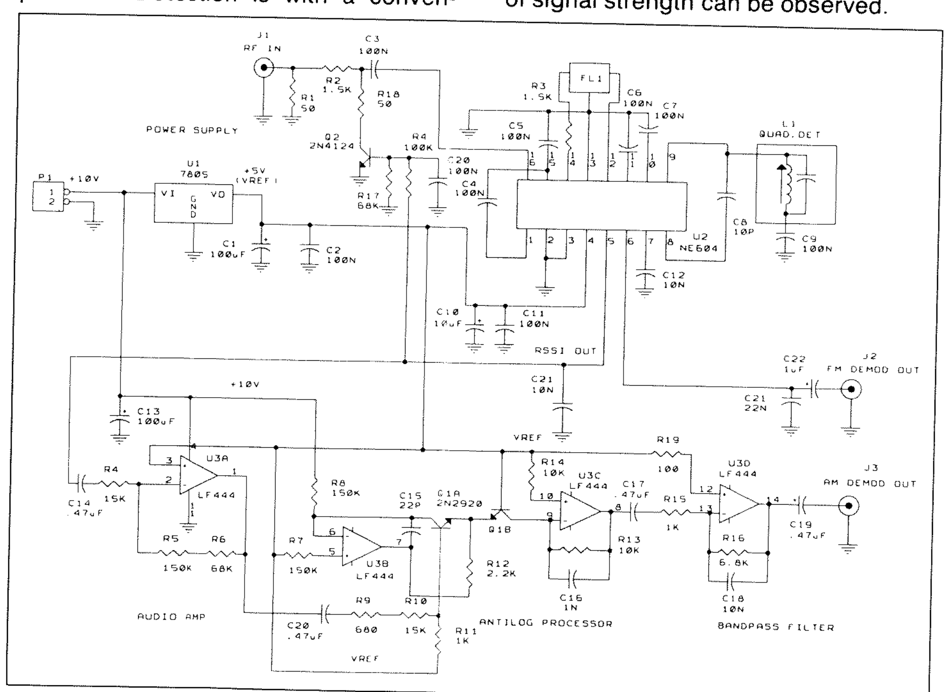

Figure 1. Schematic of AM/FM IF strip.

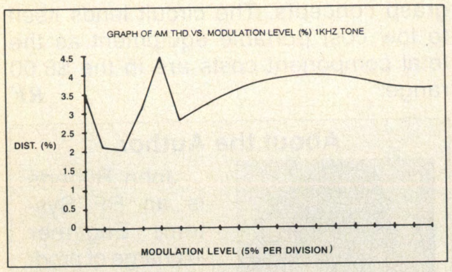

Figure 2. Distortion vs. AM modulation (%) (RFIN -70 dBm)

The main point to be put forth here is that the output current (and hence the voltage) from this pin follows amplitude variations in a logarithmic manner. It follows then, that if one were to introduce a carrier whose amplitude varies at some sinusoidal rate (i.e., amplitude modulation) to the RSSI detectors, one would expect to see an output of log10 Sin(X) at the output.

However, because logs become undefined for values less than zero and negative for values less than 1, (and this clearly did not occur at the output) the actual detected output for a sine wave modulated RF input waveform is of the form [log10Sin(x) + P] where P is some positive value greater than 1. This value P is the DC offset of the RSSI detector caused by the steady state carrier power of the input RF signal.

It can also be theorized that if the recovered AM information (which is AC coupled off the RSSI pin) could be processed in an anti-log form, the original sine wave information could be recovered. This is the function of the op-amp circuitry shown in Figure 1.

Anti-Log Amplifier

Referring to the circuit schematic, the LF 444 FET input IC op-amp is used as an input amplifier (U3A), anti-log amplifier (U3B and U3C), and a bandpass filter (U3D). Q1A is driven from the signal input (the RSSI output of the SA604) via the resistive divider formed by R10 and R11, which sets the gain of the amplifier to 1 V/decade. Q1A and U3B combine to drive Q1B in proportion to the input voltage. Q1B's collector current varies exponentially with its V_{be} and U3C gives an accurate voltage representation of this action. As the two transistors depend on each other for signal processing, a matched pair was an obvious choice. The 2N2920 has two NPN transistors fabricated on the same chip, and its small footprint lends itself to miniature designs.

This antilog function will then complement the log function that is inherent in the RSSI processing of the SA604. The only problem that may manifest itself is the possibility of distortion when overmodulation occurs. In a correctly designed radio system this is not a problem, however.

Figure 2 shows AM distortion as a function of modulation depth. It can be seen for a large modulation index in a mobile communications environment, overall distortion is comparable to or better than conventional AM detection methods.

Another benefit of this system is that it is entirely possible to run the IF strip with little or no AGC! As the recovered modulation level is held constant regardless of input signal level, (the DC level at the RSSI pin will change, but not the AC superimposed modulation) no

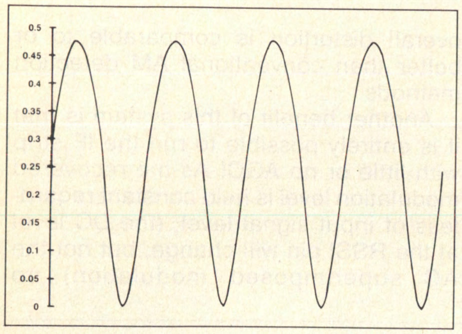

Figure 3a. Recovered AM modulation (RSSI out) SA604 Pin 5.

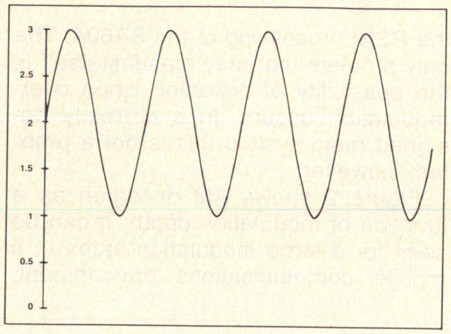

Figure 3b. Recovered AM modulation after antilog processing

method of gain control is needed over a 70 dB dynamic range! If higher dynamic range is needed, it is a simple matter to introduce some form of RF AGC at the input. The circuit is not at all temperature sensitive as the signals which it processes and recovers are all AC coupled, hence immune to DC offset errors associated with log/antilog amplifiers. There are no adjustments needed in the AM section, and it allows the use of the existing FM IF filters and associated support components.

Figures 3a and 3b show the signal recovered from the RSSI pin before anti-log processing, and the recovered waveform after processing.

Conclusion

The circuit described here presents itself to be very versatile as a multi-mode modulation detection circuit. It has proven itself to be very easy to get up and running, is stable, requires no external AGC loop, is very low cost and can be analyzed with few difficult to grasp concepts. The circuit lends itself to low cost portable equipment as the total component costs are in the $6.00 range. ***RF***

About the Author

John Roberts is an RF Systems Engineer in charge of product development at Northern Airborne Technology Ltd. He is currently designing a series of multi-mode, fiber-optic controlled VHF/UHF transceivers. His interests range from design of extremely high dynamic range receivers to low cost multi-mode portable gear. John has also designed HF systems for both commercial and government use. He can be reached at Suite 14, 1925 Kirschner Road, Kelowna, British Columbia, Canada V1Y 4N7. Tel: (604) 763-2232. Fax: (604) 762-3374.

High Resolution Plotting Routines for BASIC and APL

By Douglas Miron, Ph.D.
South Dakota State University

Engineers who write their own computation programs want simple interfaces, that are easy to use and do not require lots of time to write. The plotting routines presented here are not for professional programmers or application writers. They are intended to be subroutines or post-processing accessory programs for engineers who use the computer as a problem-solving tool, for the display of filter passbands, amplifier bandwidth, antenna pattern rectangular plots, or other typical RF engineering data.

Two implementations of the plotting routine were developed; one in Microsoft's GWBASIC, the other in STSC's APL*PLUS/PC. The BASIC version was created because of the widespread use of this language. The APL version was developed because it is the author's most often used language.

Program Description

This plotting program produces scales and a grid, with increments in nice, round numbers and labels for each axis. The plot occupies as much of the screen as possible for best resolution. The program structure is as follows:

First, maximum and minimum data values are requested. These can be expected values, or a sorting routine can be included in the user's program to identify these values if the data are computed before plotting. Next, scale extremes and major divisions are computed. Scale factors to match the pixel-to-datum value are implemented. Grid lines are then drawn, adjusted to fit round number major divisions. Scale numbers and labels are placed on the plot.

Finally, the line-drawing function is implemented. Data at each point on the horizontal scale has the corresponding vertical scale value computed and written to the screen, using both pixel and scale factors already established. The

```
10 'Some generally useful constant.
20 KEY OFF:WIDTH 40:CLS:LGT=LOG(10):PI=4*ATN(1):EPS=.5-.000001:RD=PI/180
30 'LINPLOT, a general-purpose plotting program written by D. B. Miron,
40 'March, 1990.
50 'This program is set up to do a plot while the data is being calculated,
60 'rather than after all the data has been generated.  The user thus has a
70 'sense of progress, which is encouraging when the computation for each
80 'point takes a long time.  To achieve this effect, the scales, labels,
90 'and plotting are the main program, and the data function is a subroutine.
100 'There is a demonstration function at the end of this program.  Replace it
110 'with yours.
120 'While the input-request format is good for a single use, it becomes tedious
130 'during problem-solving in which many parameter values are to be tested, or
140 'many changes in the math model are to be tried.  I suggest that you comment
150 'out the INPUT statements and type in a line assigning the scale values and
160 'labels you will be using.
170 'If a data value is out of range for the scales you chose, the program
180 'offset it by a multiple of the range so that it appears on screen.  This
190 'process will cause a jump in the plot, which tells you  there's an
200 'error somewhere.  I prefer an incorrect curve to a blank screen when I
210 'a mistake.
220 'To get a printer copy of your plot, you must run DOS's GRAPHICS.COM before
230 'you start BASIC.  The plot is done in 640x200 CGA mode, so GRAPHICS prints
240 'it siedeways.  It takes a full sheet, so be sure the printer is set to
250 'start at the top.
260 LOCATE 1,1,0
270 INPUT "Min. and max. horizontal values";XMN,XMX:CLS
280 LOCATE 1,1,0:INPUT "Horizontal units";UNX$:CLS
290 LOCATE 1,1,0
300 INPUT "Min. and max. vertical values";YMN,YMX
310 CLS:LOCATE 1,1,0:INPUT "Vertical units";UNY$
320 CLS:LOCATE 1,1,0:INPUT "Caption (80 characters max.)";CAP$
330 PXX=10^INT(LOG(ABS(XMX-.000001))/LGT):PXN=10^INT(LOG(ABS(XMN+.000001))/LGT)
340 IF PXX>PXN THEN PX=PXX ELSE PX=PXN
350 RXX=CINT(EPS+XMX/PX):RXN=INT(XMN/PX)
360 NX=RXX-RXN:IF NX>10 THEN PX=PX*10:GOTO 350
370 IF NX<3 THEN PX=PX/5:GOTO 350
380 RX=NX*PX:XSF=400/RX:DX=(XMX-XMN)/200:XPS=120:YPS=170
390 PYX=10^INT(LOG(ABS(YMX-.000001))/LGT):PYN=10^INT(LOG(ABS(YMN+.000001))/LGT)
400 IF PYX>PYN THEN PY=PYX ELSE PY=PYN
410 RYX=CINT(EPS+YMX/PY):RYN=INT(YMN/PY)
420 NY=RYX-RYN:IF NY>10 THEN PY=PY*10:GOTO 410
430 IF NY<3 THEN PY=PY/5:GOTO 410
440 RY=NY*PY:YSF=YPS/RY
450 XPN=RXN*PX:YPN=RYN*PY:YPX=RYX*PY:SCREEN 2,0
460 FOR K=0 TO NY:FOR X=XPS TO 520 STEP 4:PSET (X,K*PY*YSF):NEXT X,K
470 FOR K=0 TO NX:FOR Y=0 TO YPS STEP 2:PSET ((XPS+K*PX*XSF),Y):NEXT Y,K
480 FOR K=0 TO NY:LOCATE 1+K*PY*YSF/8,((XPS/8)-3-CINT(ABS(LOG(PY)/LGT)))
490 PRINT (RYX-K)*PY;:K=K-(NY>5):NEXT K
500 FOR K=0 TO NX:LOCATE 1.5+YPS/8,(XPS/8)+K*PX*XSF/8
510 PRINT (RXN+K)*PX;:K=K-(NX>5):NEXT K
520 LOCATE 2,1,0:PRINT UNY$;
530 LOCATE 1.5+YPS/8,((XPS+424)/8+CINT(ABS(LOG(ABS(RX))/LGT))),0:PRINT UNX$;
540 LOCATE 25,1,0:PRINT CAP$;
550 X=XMN:GOSUB 660:GOSUB 640
560 PSET ((XPS+(XMN-XPN)*XSF),(YPS-(Y-YPN)*YSF))
570 FOR K=1 TO 200
580 X=XMN+K*DX:GOSUB 660:GOSUB 640
590 LINE -((XPS+(X-XPN)*XSF),(YPS-(Y-YPN)*YSF))
600 NEXT K
610 KB$=INKEY$:IF KB$="" GOTO 610
620 SCREEN 0,0
630 END
640 IF Y=<YPX THEN GOTO 650 ELSE Y=Y-RY:GOTO 640
650 IF Y>=YPN THEN RETURN ELSE Y=Y+RY:GOTO 650
660 Y=2*X/(5*(1+X*X/25)):RETURN
610 KB$=INKEY$:IF KB$="" GOTO 610
620 SCREEN 0,0
630 END
640 IF Y=<YPX THEN GOTO 650 ELSE Y=Y-RY:GOTO 640
650 IF Y>=YPN THEN RETURN ELSE Y=Y+RY:GOTO 650
660 Y=2*X/(5*(1+X*X/25)):RETURN
```

Figure 1. Listing of the BASIC plotting routine.

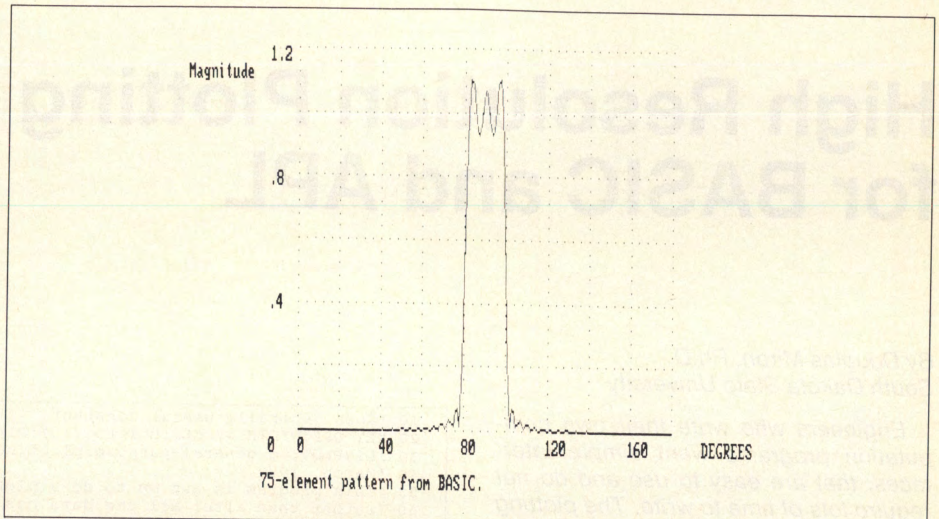

Figure 2. Antenna pattern plot example (BASIC version).

program waits for a keyboard input to exit. The user could include a simple "beep" or other command to inform him or her that the plot is finished.

The BASIC version uses the 640×200 pixel CGA graphics mode, with 201 data points on the horizontal axis for a smooth curve. The operation is "compute as you plot," so the curve crawls across the screen as it is written. It can easily be modified to use data already computed. A listing of this version is shown in Figure 1.

APL may not be familiar to the reader,

and this is probably not the place for a tutorial on APL use. Engineers familiar with APL know that it is an excellent array processing language. A variable name can stand for a scalar, a vector, or a multidimensional array, as well as a number or character. It has operators for arithmetic and transcendental functions, and many operators to manipulate, sort, combine and separate arrays.

User-defined functions can be used just like the built-in functions, so several have been created for using the plotting routine. CGRAF sets up the CGA graphics mode, CLEAN saves the current screen and initiates graphics parameters, CLOSE does some housekeeping and returns to the previous screen, and PGRAF sets graphics parameters for output to the printer. PLOT is the main plotting program. The PLOT function works in the same general way as the BASIC version, but it assumes that data have already been calculated. PSET is a cover function similar to the BASIC command, but more powerful, and VERSUS is another cover function that formats X and Y values so they will be connected with lines.

Example

Figure 2 shows a plot using the BASIC version of an antenna array of 75 elements, fed with currents having even symmetry about the center element. The broadside pattern function is:

$$f(\theta) = C_o + \sum_{k=1}^{m} C_k \cos(\beta d \cos \theta) \qquad (1)$$

in which,

θ is the direction angle measured from the array axis.

C_k are proportional to the currents

β is $2\pi/\lambda$

λ is the wavelength

d is the element spacing

A BASIC subroutine executing equation (1) was written to generate the plot. Although not shown, the APL version has better resolution using the same CGA mode, due to better graphics capability.

Conclusion

If the user wants flexible plotting capability for BASIC or APL engineering programs, these routines can make the job easy. Both versions, along with additional variations, are provided on disk in this month's offering from the RF Design Software Service. See page 8 for ordering information. **RF**

About the Author

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The Maturing of High Performance Software

By Liane G. Pomfret
Associate Editor

High performance software has undergone a rapid evolution in the last few years. In many cases, the software packages of five years ago bear little resemblance to their offspring of today. Advances in computing hardware, programming and algorithms have changed the entire CAE industry.

Fifteen years ago, software for designing circuits was extremely expensive, cumbersome, and usually required a mainframe to run on. Since then, software packages have advanced to where they are now capable of running on PCs and workstations, can do both linear and nonlinear simulation, and are user friendly. Ray Pengelly, Vice President of Marketing and Sales for Compact Software says of their software, "The product in all forms, whether it be for PCs, workstations or mainframes, is very different from where it was even two years ago."

Virtually all of the large software companies are pursuing the goal of operating their software products within a major integrated design framework. Jeremy Bunting, product marketing manager at EEsof indicates that EEsof is doing this with "Mentor and Cadence, and looking to establish programs with Integraph, ComputerVision, Valid, and others — We're incorporating our product suite into their frameworks." This way engineers can do everything from initial design to circuit or system simulation without ever leaving their desk. "Before frameworks, the design process was cumbersome due to individual point tool characteristics," stated Paul Wilhelmson, channel marketing manager for Cadence. He continues, "Now designs can be created efficiently because tools are integrated together under a common user interface. This allows engineers to create design flows matching their design methodologies."

The biggest developments have been in the areas of nonlinear simulation, system simulation, user defined models, component libraries, and schematic capture and layout. John Hirsekorn, Product Marketing Manager for Hewlett-Packard's HF Design Software notes, that, "We're improving simulation accuracy, but even more profound is our ability to model high frequency transmission media, passive components, and perhaps most importantly, nonlinear characteristics of active devices." In addition, there have been significant advances which make the software more user friendly and consequently more marketable. Randy Rhea, president of Eagleware comments, "a wider range of types of circuits can now be successfully simulated. There are more models to be used within these circuits and the products have also gotten much easier to use."

Electromagnetic Simulation Software

One of the largest areas of software development is electromagnetic simulation. The need for software in this area has been acute and companies are just beginning to develop it commercially. According to James Rautio, president of Sonnet Software, "Most electromagnetic software, except for the past two to three years, has been academic or research oriented. Software support and updates are minimal and the software is usually difficult to use." The new software packages offered by companies such as Sonnet, Hewlett Packard, EEsof and Compact provide engineers with a way to see what is going on inside the circuit, electromagnetically. With this software, "Customers can achieve packing densities roughly two to three times what they were able to do before," says Rautio. It can also cut down on the length of the design cycle and significantly lower costs.

PC Versus Workstation

Today's engineer has a number of choices when it comes to computing hardware. The mainframe computer is still an alternative but is giving way to smaller, less expensive options such as PC compatibles, workstations and Macintosh computers. Each system has its supporters and all have their pros and cons. The Macintosh versus PC compatible debate has been going on for several years. According to Brad Nedrud, Owner of Nedrud Systems, "We're running into more and more engineers that have Macs and like using them." Zelco Jagaric, manager of technical support and development at Ingsoft notes that, "The only platform which actually delivers high performance general purpose applications in all fields, completely standardized and fully integrated, is the Macintosh." Indications from both these companies are that there is a growing need for Mac CAD software, despite assertions from other companies that there is no market.

With the power of PCs approaching that of the lower end workstations, engineers now have to decide between a dedicated system, i.e. a workstation for just CAD work, or a general purpose tool, i.e. a PC that can also do spreadsheets, etc. DGS Associates president, George Szentirmai, has noticed a trend in his sales, "Lately the PC version has become the most popular form. It is the easiest to use because of utilities that are not standard on mainframes and workstations." Dick Webb, president of Webb Laboratories comments "Our network analyzer will run five times faster on a 486-based PC, than a similar program on a low end workstation."

Supporters of workstations, some of which now approach the power of a mini-mainframe, feel that they are better suited to getting tasks done. As Jeremy Bunting points out, "With its ability to network, its high resolution graphics and great computing power, the workstation becomes a very powerful tool."

The choices available in high performance software are staggering. But the decision must also take into consideration the platform the software will run on, the number of users, cost, the frequency of use and applications. It may be that one framework will provide all the answers an engineer is looking for but another engineer may find different tools from several companies that are right for him. Either way, a great deal of time should be spent researching and testing potential software packages before making a final decision. **RF**

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RF software

Electromagnetic Simulator

Sonnet Software has announced an updated release of em™ 2.2, electromagnetic analysis software. Em is a full-wave 3-D electromagnetic simulator optimized for predominantly planar circuits such as microstrip and coplanar waveguide. The analysis now includes "balanced" or "push-pull" ports, an automatic box resonance identification and/or removal option, separate minimum and maximum subsection size, and the ability to output a "jxy" file. The user interface now allows direct input from GDSII Stream and from Cadence's Analog Artist™.
Sonnet Software, Inc.
INFO/CARD #210

EMI/EMC Testing

The Commercial EMI Test Software Package from Tektronix allows product designers and manufacturing QC people to prequalify electronic products for new and tougher EMC requirements. The software runs on XT-class and higher personal computers and has the following capabilities: radiated emissions tests, VDE magnetic emissions tests, CISPR/VDE conducted emissions tests and others.
Tektronix, Inc.
INFO/CARD #209

Radio System Design

The Radio System Design Software from JMA Applied Spectrum Research is now available. The RSDS software allows for both single and multi-site design, frequency re-use planning, "real world" propagation studies based on digital topographic terrain and FCC broadcast and Carey contours. The system can help plan all mobile radio bands, including trunked radio, European cellular and the new 200 MHz service.
JMA Applied Spectrum Research
INFO/CARD #208

User Interface Development

Hewlett-Packard has announced HP BASIC Plus, a software package that simplifies user-interface development. The software reduces the number of lines of programming code required to create user interfaces and runs on all HP 9000 series 300 controllers. HP BASIC 6.2 is required to run HP BASIC plus. Cost for the HP BASIC Plus and HP BASIC 6.2 is $450 and $1050 respectively.
Hewlett-Packard Company
INFO/CARD #207

Models for HSPICE

Meta-Software is now offering TriQuint GaAs modeling capabilities in its HSPICE circuit simulation software. TriQuint's Own Model assures more accurate GaAs modeling using HSPICE by correcting deficiencies in existing SPICE models while adding several features unique to GaAs. The software is available on PCs, workstations and mainframes.
Meta-Software, Inc.
INFO/CARD #206

Testing Software

AutoCAT automates production or laboratory tests by interfacing with measuring instruments, through IEEE-488 or RS-232 ports, and controls these instruments to manufacturer's specifications. Some of its features include: user controlled data format, spreadsheet compatible data files, add custom instruments to its library, read, average and stream data from instruments. The software requires an IBM PC or compatible, 640 K RAM, 1 floppy disk drive and a graphics card.
NEOS
INFO/CARD #205

Arbitrary Waveform Generator Tool Kit

Wavetek has introduced a DSP software tool kit for synthesizing, designing and analyzing analog and digital waveform signals for an arbitrary waveform generator. For use on a PC, WaveForm DSP can capture, create, edit or analyze waveshapes on the PC screen and then upload digitized waveshapes to an arbitrary waveform generator. Additional features include a standard library of more than 20 waveforms that can be used in creating or editing waveforms. The user can also add custom waveforms to the library for reuse. Pricing is $895 with delivery at 4-6 weeks.
Wavetek Corporation
INFO/CARD #204

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Special Courses

OCTOBER 28, 29, AND 30

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

Fundamentals of RF Circuit Design: Part I — Oct. 28

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 — Oct. 29

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 of bias, temperature, parasitics and losses are considered, and an introduction to tolerance analysis is presented. Instructor: Les Besser, president, Besser Associates Inc.

Filter and Matching Network Design: L-C and Distributed Circuits—HF to Microwaves — Oct. 28

This course is designed for the practical engineer, packing a wealth of practical and useful information on these passive RF circuits into eight hours. Engineers with all levels of experience will benefit from the review of fundamental information on filter response and classic topologies for filters and matching networks, followed by design methods for implementation of L-C and distributed-element filters and matching networks, from low radio frequencies to microwaves. key performance parameters such as group delay and phase characteristics are covered, as are techniques for implementing the design of these networks using modern computer-aided synthesis, analysis and optimization. Instructor: Randy Rhea, Eagleware/Circuit Busters.

Oscillator Design Principles — Oct. 30

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.

Ham Radio Reception: "The Magic of RF"

A standing-room-only favorite at the 1990 RF Expo East, the Amateur Ham Radio Reception will be held again in 1991. One of the highlights of this event will be the drawings for outstanding door prizes, and drinks and munchies will be provided. Attendees are encouraged to bring their QSL cards for posting on the bulletin board and plenty of tall tales to swap with fellow hams.

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

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Who Should Attend

RF Expo East is attended by electronics engineers and engineering managers working in the RF frequencies. Engineers new to high frequency circuits benefit from the numerous fundamental and tutorial programs offered. Senior engineers receive valuable updates on the latest developments in high frequency technology. In 1990, attendance was as follows:

Engineers	53%
Engineering Managers	15%
Engineering Service Professionals	7%
Owners or Officers	15%
Others (including reps, buyers, consultants)	10%

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Englewood, CO 80111

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RF literature

Fast Filter Discriminators

K&L Microwave has released a new data sheet for their modular fast filter discriminators. The sheet contains information on operation characteristics, including charts, architecture, and standard features. Specifications and performance characteristics are given for discriminators in different frequency ranges.
K&L Microwave, Inc.
INFO/CARD #200

Circuit Fabrication Design Guide

Poly Circuits has a new brochure where the costs of PTFE microstrip and stripline circuits can be engineered by the customer. Outlined are the cost factors for etch tolerance, dimensional tolerance, plated hole tolerance, front to back registration, line width, aspect ratio, pad size large than hole, and metalization. Further information on any process will also be made available upon request.
Poly Circuits, Inc.
INFO/CARD #199

Coaxial Connectors

A catalog featuring a line of RF coaxial connectors, waveguides and related components is being offered by Tru-Connector. Catalog D features pulse series ceramic, BNC, C, SM, N, HN, LC, LN, small twin series weatherproof, high-voltage miniature, QDS and QDL, BN and UHF connectors. Miscellaneous connectors, parts, cable terminations, adapters, waveguide to Type N adapters, waveguide to coax adapters and submersible cable to cable connectors are also included.
Tru-Connector Corporation
INFO/CARD #198

RF & Cellular Components

M/A-COM has released a new catalog of Anzac RF & cellular components. The featured product lines include attenuators and switches, I/Q modulators, power splitters/combiners, and mixers. Part numbers, specifications and pricing for each item is given as well as dimensions and materials.
M/A-COM, Anzac Operations
INFO/CARD #197

Log Amplifiers

Data sheets describing Akon's line of IF Log Amplifiers and Detector Log Video amplifiers are now available. The units operate from 20 MHz through 20 GHz in IF, Radar, Communications and ECM/EW bands. The data sheets give examples of products available, with custom designs available upon request.
Akon
INFO/CARD #196

High Speed Linear Products

A new brochure features high-resolution, high-speed A/D and D/A converters, sample/holds, operational amplifiers and analog multipliers used in instrumentation, imaging, video, spectrum analysis and direct digital synthesis. 'High-Speed Linear Products' from Burr-Brown gives a quick summary of available products including key specifications and application ideas.
Burr-Brown Corporation
INFO/CARD #195

IF/RF Components

RF Components has released its new IF/RF Components Catalog. The catalog details their line of log amplifiers, limiting amplifiers, frequency discriminator, and phase detectors. Product data sheets within the catalog include guaranteed specifications and typical performance as well as unit pricing. The catalog also includes some brief application notes for their log amps.
RF Components, A Subsidiary of Signal Technology Corporation
INFO/CARD #194

Between Series Adapters

A 24-page brochure from ITT Cannon/Sealectro details adapters that are widely used for high efficiency transitions between all series of RF connectors. The brochure has a reference chart to locate the proper adapter at a glance. Specifications are listed, including operating range and insertion loss, as well as detailed line drawings with dimensions.
ITT Cannon Sealectro
INFO/CARD #193

Hybrid Clock Oscillators

Spectrum Technology has introduced a new 1991 catalog describing their range of hybrid clock oscillators, TCXOs and TCVCXOs. The catalog lists all the necessary features of their numerous standard oscillators. Clock oscillators are available in a variety of packages for the commercial, military and aerospace markets.
Spectrum Technology, Inc.
INFO/CARD #192

RF Connectors

The latest catalog from Amphenol RFX details their line of BNC, TNC, Type N, UHF, Twinax, Coaxial and Twinaxial connectors for commercial applications. The catalog contains indices, cross references, specifications, line drawings with dimensions, assembly instructions, and actions cards for engineers requiring more information.
Amphenol Corporation
INFO/CARD #191

Universal Test Fixture

A new 8-page brochure describes a new universal test fixture for substrate measurements. Detailed information on microstrip, coplanar waveguide, offset, and right angle measurements, all without the need for custom carriers is included. Complete specifications and ordering information are included.
Wiltron
INFO/CARD #190

RF design

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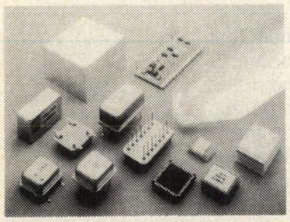

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Microwave Circuit Analysis

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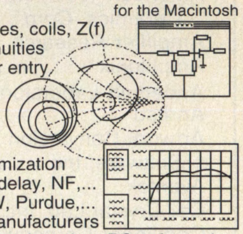

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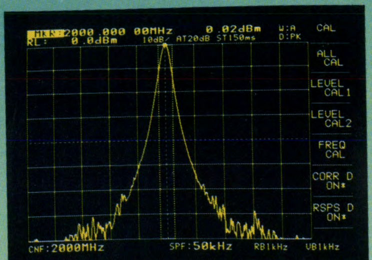
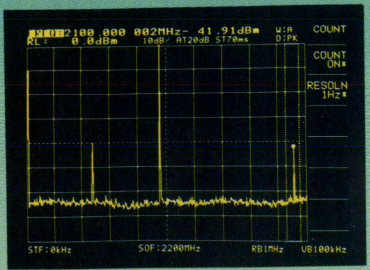
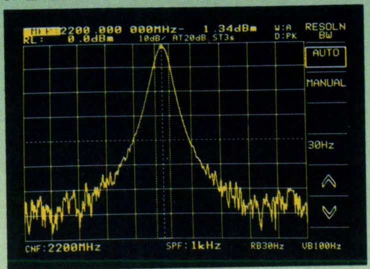


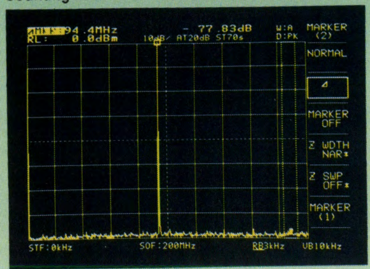
Overall Accuracy Level of ±1dB



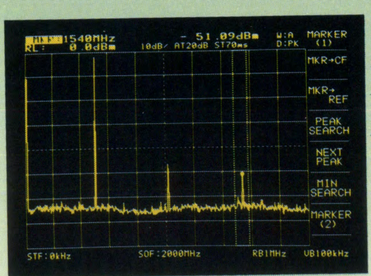
Automatic Tuned Frequency Counting with 1Hz Resolution



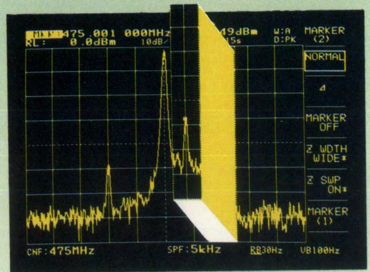
30 Hz Resolution Bandwidth



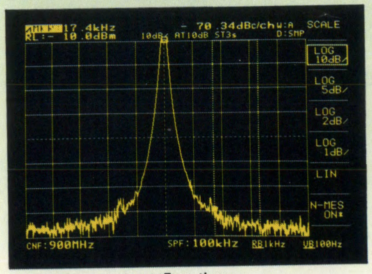
75 dB Dynamic Range



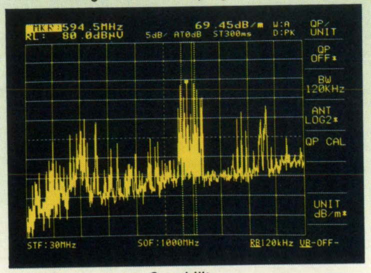
Signal Capturing Zone Marker



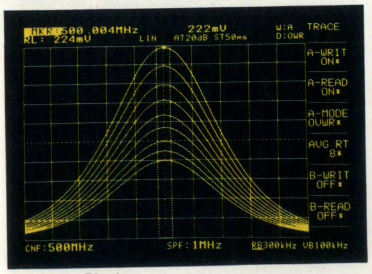
Reduction of Measurement Time Through Zone Sweeping



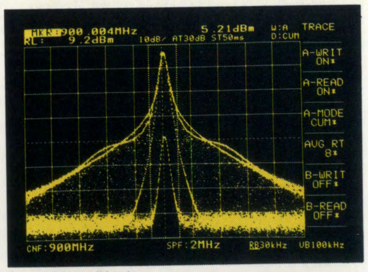
Noise Measurement Functions



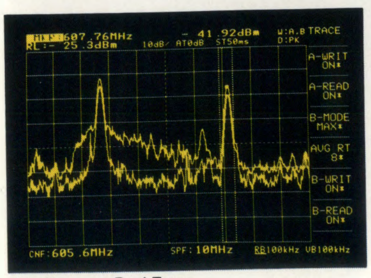
EMI Measurement Capability



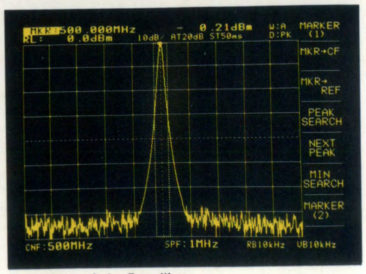
Overwrite Display



Cumulative Display



Simultaneous Dual-Trace Display



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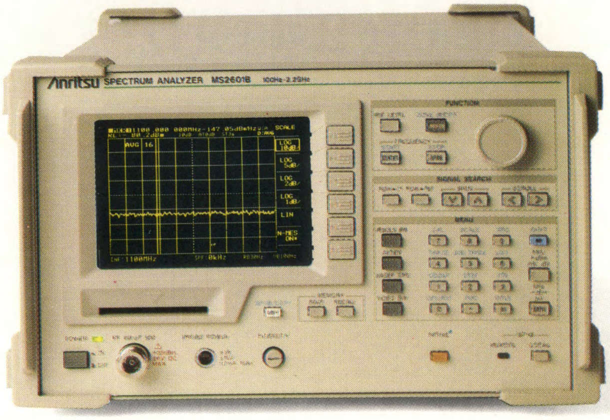

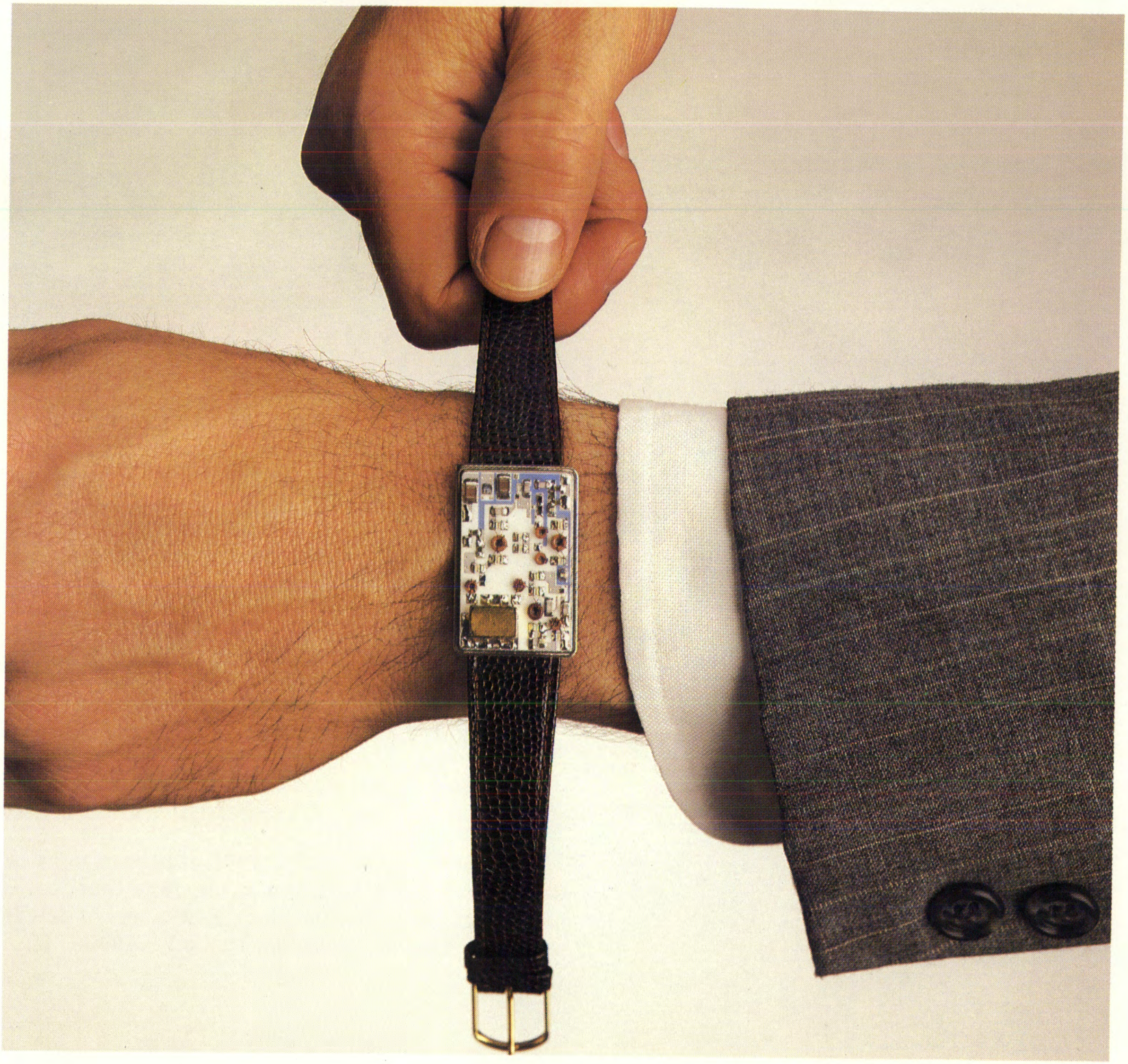

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