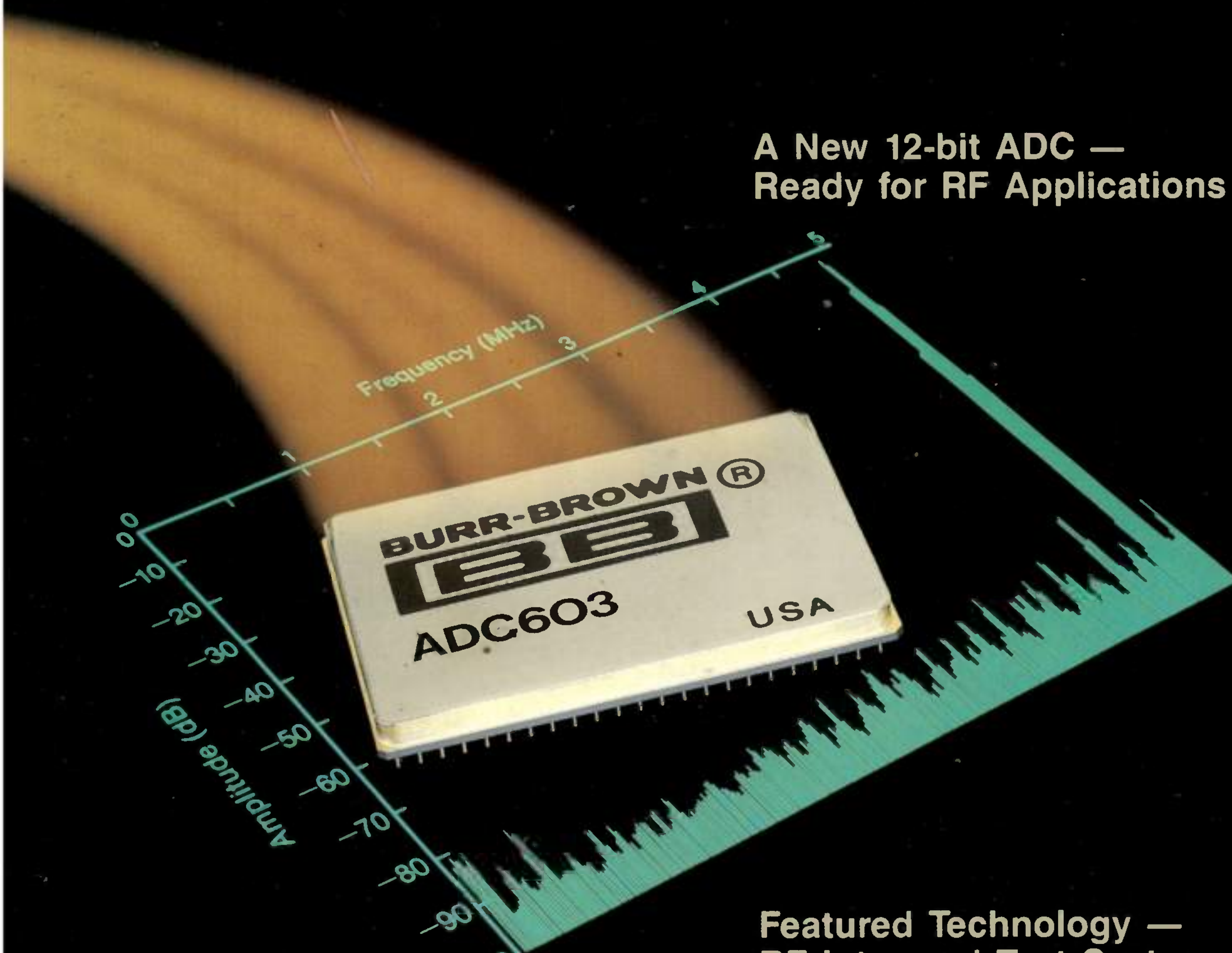
rf design

high frequency and analog engineering

November 1987



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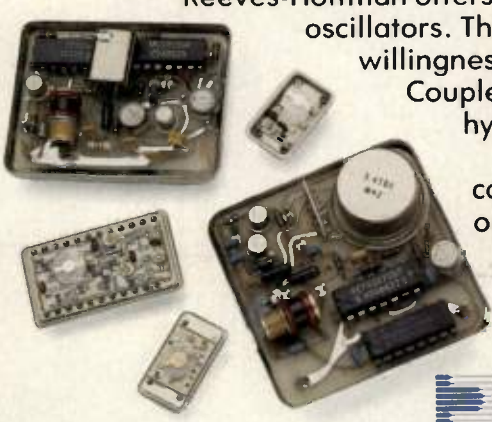

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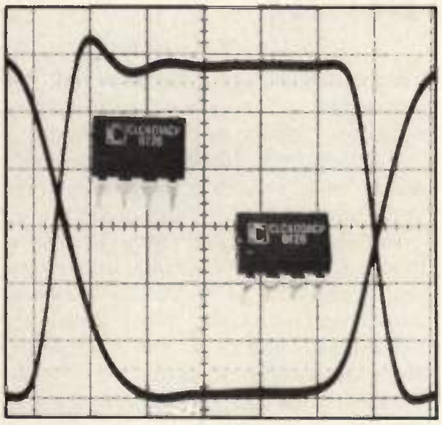

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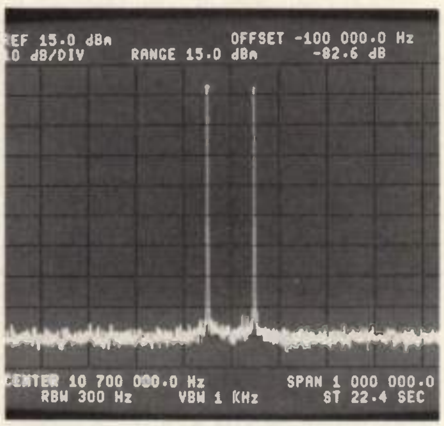

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rf editorial

RF: A New Definition

By Gary A. Breed
Editor

RF (ar' ef) */radio frequency/,* n. 1. that portion of the electromagnetic spectrum useful for radio transmission, lying (roughly) between audio and infrared. 2. electronic signals operating within this frequency range or at analogous speeds.

The distinguished history of radio transmission is reflected in the traditional definition of RF as a segment of the electromagnetic spectrum (1.). This can be considered the "pure" definition, based on the original concept of radio as communication using electromagnetic waves.

Most dictionaries also include a second definition to account for signals that are contained within circuits, typically using the terms "amplification" and "detection" to identify their functions. The first half of (2.) in our new definition retains this secondary meaning of RF.

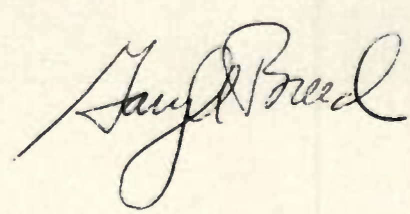
All the definitions I have seen stop at this point, which seems to limit the meaning of RF to those things which ultimately involve electromagnetic waves — radio equipment. Certainly, when RF was first defined, only radio circuits operated at these frequencies. However, present electronic technology offers too many contradictions if we accept such a limitation.

By adding the phrase ". . .or at analogous speeds," we have given RF a more universal meaning, in keeping with the advances in technology. There are many digital and analog applications that are not radio signals, yet they are characterized by MHz and GHz frequencies and nano- and picosecond operation times. These applications require engineers to understand "radio" techniques: component behavior at high frequencies, lumped and distributed circuit elements, transmission and reflection, bandwidth, phase accuracy, linearity, and noise. These circuits can also unintentionally transmit and receive electromagnetic waves, which is reason enough to give them attention!

The commonality is quite clear when we consider that pulse modulation can drive either a radar system or a fiber optic laser diode, ECL logic devices can be used as mixers and amplifiers as well for digital applications, transistors from the same family are used as high-resolution CRT drivers as well as RF power amplifiers, and an FM receiver IC can demodulate either audio from a radio system or data from a local area network. As further evidence, manufacturers of high speed operational amplifiers (usually considered "non-RF" devices) have begun to characterize their S-parameters.

There is plenty for the "radio" engineer to learn from his high speed analog and digital counterparts: time domain analysis, information theory, DC-coupled circuits and control theory. Sharing ideas has other benefits, too. A fascinating part of today's engineering is finding applications for components that their manufacturers never intended, such as using video buffers as RF drivers, or consumer radio ICs in test instruments.

The new definition of RF has us excited! The technology it defines is rich with possibilities, and *RF Design* will the be the only place where these universal high frequency/high speed design techniques come together.



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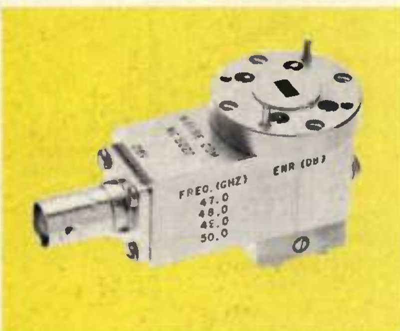

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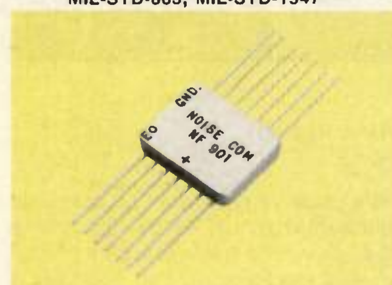

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rf viewpoint

Go for the Gold

By Keith Aldrich
Publisher

In this issue you'll find an ad announcing the third annual RF Design Awards contest, complete with news of the prizes we've arranged through the good will of some of our fine advertisers.

This contest has come to be a solid institution in the short time since we started it, and to be one of the winners is possibly the leading recognition that can be earned for RF engineering ingenuity. The grand prize, whose winner is the subject of our cover illustration in the July issue, is the RF community's equivalent of an olympic gold medal.

Why not "go for the gold" in 1988?

We received just thirty entries in our 1987 contest (won by Charles Wenzel of Wenzel Associates, Austin, Texas), but every one of them was the fruit of many hours of labor. It is not an idle decision to be a participant in the contest because, first, of the time commitment, and second, because of the challenge it will pose to your gifts. A routine solution to a routine circuit problem obviously isn't going to earn you the admiring applause of your peers. You are going to have to present a really novel yet eminently practical circuit, solving some problem familiar to RF engineers in a way that is not so familiar — the kind of thing that makes people snap their fingers and say, "Why didn't I think of that?"

This emphasis on new ways to solve familiar problems is reflected in the choice of the grand prize and the runner-up prize.

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The chance to compete for such lavish yet useful prizes should be a tremendous incentive for many of you to enter the 1988 RF Design Awards contest. . .yet they probably will not be the main incentive to those of you who are truly competitive and like to excel. The main incentive will be the chance to "show your stuff," and demonstrate that it is the stuff of which champions are made.

There aren't enough occasions in our society to celebrate winners in contests other than physical ones. Don't let this rare opportunity to shine get by you.

Go for the gold!

Keith Aldrich

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Keith Aldrich

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Gary A. Breed

Assistant Editor
Mark Gomez

Consulting Editors
Andy Przedpelski
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Sales Supervisor
Kate Walsh

Advertising:
Western States
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36 Belmont Rd. S.W. 3
West Harwich, MA 02671
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SP2T	5-200	DS0492	0.5 / 0.8	52 / 40	20.0	TTL	+5, −15 / 30	+30	14 PIN DIP
SP2T	20-2000	DS0612	1.6 / 2.3	45 / 50	10.0	TTL	+ 5 / 3	+10	14 PIN DIP
SP2T	200-900	DS0432	0.5 / 0.8	45 / 35	1.5	TTL	+5, −15 / 30	+35	14 PIN DIP
SP2T	500-2000	DS0257	0.8 / 1.5	40 / 35	0.4	TTL	± 5 / 25	+10	14 PIN DIP
SP3T	50-500	DS0073	0.8 / 1.0	45 / 40	2.0	CMOS	+12 / 5	+10	0.8 sq 24 PIN
SP4T	10-500	DS0085	0.6 / 0.8	40 / 25	35.0	CMOS	+15 / 6	+10	14 PIN DIP
SP4T	500-2000	DS0259	1.3 / 1.5	50 / 25	0.4	TTL	± 5 / 35	+10	24 PIN DIP
SP5T	200-800	DS0475	0.6 / 1.0	50 / 40	1.0	TTL	± 5 / 27	+10	24 PIN DIP
SP6T	100-600	DS0506	0.6 / 1.0	56 / 50	1.0	TTL	± 5 / 40	+20	38 PIN DIP
SP8T	100-1100	DS0518	1.1 / 1.6	45 / 35	1.0	TTL	± 5 / 45	+13	24 PIN DIP
XFER	15-45	DS0097	.5 / 1.0	60 / 50	10.0	CMOS	+ 5 / 20	+10	0.8 sq 24 PIN
XFER	70-1000	DS0319	1.25 / 1.5	60 / 40	10.0	CMOS	+15 / 15	+10	18 PIN DIP
SPST	20-600	100C1281	1.2 / 1.3	80 / 70	0.5	TTL	+ 5 / 75	+10	SMA CONN.
SP2T	20-600	100C1282	.8 / 1.2	80 / 70	0.5	TTL	+ 5 / 110	+10	SMA CONN.
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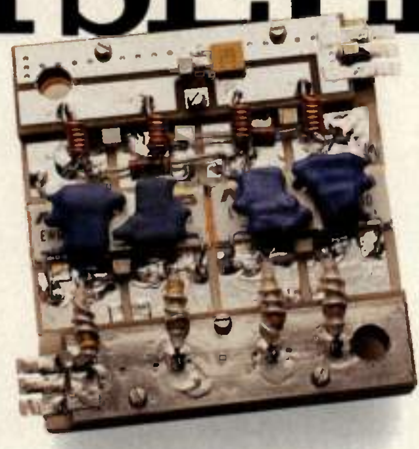

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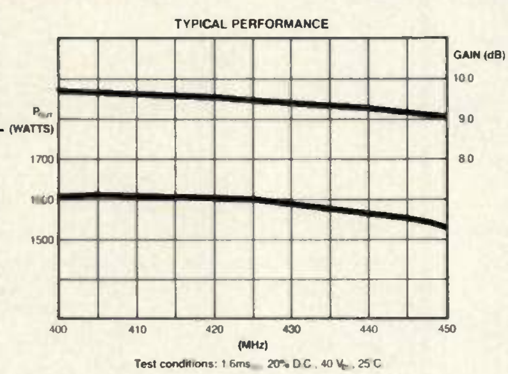

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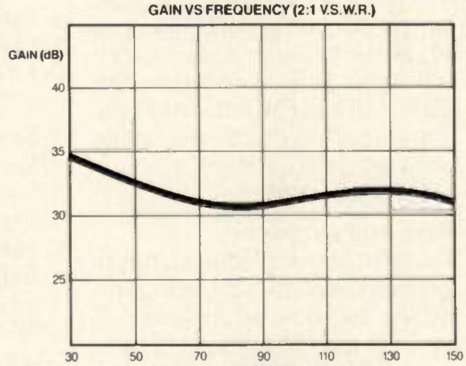

Test conditions: R [illegible] 2:1 V.S.W.R., V [illegible] 28V Supply, 10V Bias, P [illegible] 1000W Forward Delivered Power

The amplifier performs from 30–150 MHz with a minimum of 30 db gain. Its design, using power FET's, insures high gain and low noise figure.

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rf letters

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

Another Method of Phase Noise Calibration

Editor:

Subject: August 1987 issue article "Phase Noise Calibrator."

There is another mechanism of significant usefulness for providing calibration for phase noise measurement, the Armstrong Modulator (1). Figure 1 illustrates the arrangement:

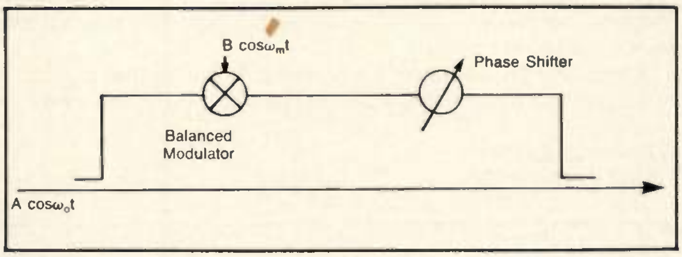


Figure 1

Figure 2 illustrates the vector relationships:

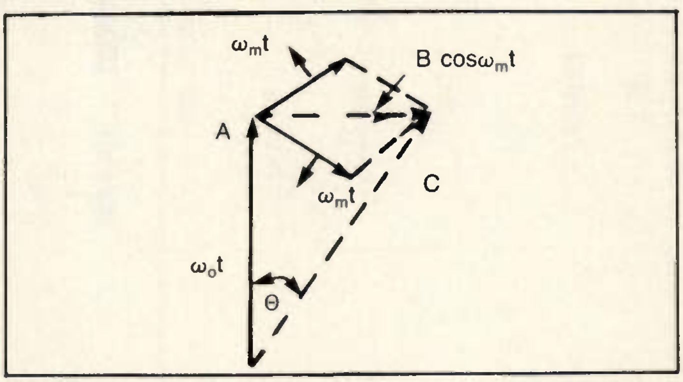


Figure 2

From Figure 2 one can find:

$$\Theta = \tan^{-1}\frac{B}{A}\cos\omega_m t$$

$$= \frac{B}{A}\cos\omega_m t - \frac{1}{3}\left(\frac{B}{A}\cos\omega_m t\right)^3 + \ldots \qquad (1)$$

$$= \frac{B}{A}\cos\omega_m t\left[1 + \frac{1}{4}\left(\frac{B}{A}\right)^2\right] + \text{harmonics}$$

For $\frac{B}{A} << 1$ (The usual case when calibrating for low phase noise measurements.)

$$\Theta = \frac{B}{A}\cos\omega_m t \text{ (There is <0.02 dB error for } \frac{B}{A} = 0.1\text{)} \qquad (2)$$

For $\frac{B}{A} << |$ (The usual case when calibrating for low noise measurements).

Also, the resultant magnitude is

$$C = [A^2 + (B\cos\omega_m t)^2]^{1/2}$$

$$= A\,[1 + (\frac{B}{A}\cos\omega_m t)^2]^{1/2} \qquad (3)$$

$$= A\,[1 + \tfrac{1}{2}\,(\frac{B}{A}\cos\omega_m t)^2] \text{ for } \frac{B}{A} << 1$$

$$= A \text{ within 0.02 dB for } \frac{B}{A} = 0.1 + \text{insignificant harmonics.}$$

Note that by setting the phase shifter for a net 0°, one creates AM having the same sideband magnitudes as the PM case. Since many systems also require knowledge of AM noise, there is significant advantage in using this calibrator.

By using a phase shifter capable of 360°, one can easily achieve calibration at any carrier frequency since the phase modulation index (LO) is determined by the modulation drive, not by any differential phase mechanism.

Since the balanced modulator can operate linearly with respect to the modulating signal, modulation frequency harmonics can be made very small and control of the desired modulation percentage is a simple control of the drive to the modulator.

In addition, modulation frequency and waveform is controlled by the modulation source, providing great flexibility in usage. A typical procedure is to use the tracking generator output of an HP3585 (or similar spectrum analyzer) as the modulation source, thus allowing a calibration plot over the sideband frequencies of interest.

This particular approach is used in Raytheon's line of carrier noise analyzers (2).

G.E. Romaine
Raytheon Company

References

1. L.B. Arguimbay, *Vacuum Tube Circuits,* Wiley & Sons, 1948.
2. Carrier Noise Analyzer, CNA-21 I/J; Raytheon Co., Waltham, MA.

Microstrip Program Correction

Editor:

We have found two subtle errors in our MSTRIP.BAS program published in the July 1987 issue of *RF Design.* Line 20 in the program should be as follows:

```
20 DIM B(50),L(50),F(50),Z0(50),LM(50),WM(50),W(50),C(50)
```

The corrections are:

1. Z0(50) instead of ZO(50). (zero and not a capital letter "o").
2. C(50) was omitted in the original program.

Since BASIC defaults to a dimension of 10, these errors would only show up if the user had a large enough circuit to require more than 10 of either variable.

D.R. Hertling
Georgia Institute of Technology

Broadband Power Divider Note

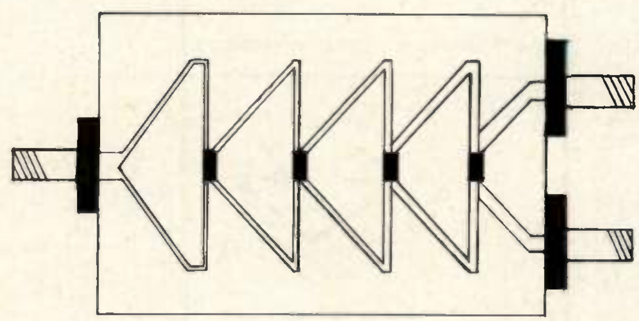
Editor:

I enjoyed Peter Vizmuller's article "Broadband Miniature Power Splitter" in the August 1987 issue. It seems to be a novel way of obtaining power division, but I am confused about the bandwidth. Mr. Vizmuller starts out by mentioning a frequency range of 500 to 2000 MHz but all of the data graphs go from 200 to 1200 MHz!

I have enclosed a design of a power divider which was done Aug. 22, 1974. This design is based on the well known article published by Seymour B. Cohn entitled "A Class of Broadband Three Port TEM-Mode Hybrids" as shown in the Feb. 1968 issue of the MTT. The figure shows a four-section divider on an alumina substrate with thick-film conductors and resistors. The printed electrical data below the figure indicates excellent performance beyond the 500 to 2000 MHz range and the size is about the same as Mr. Vizmuller's.

I thought you would be interested in knowing that the standard method sometimes works as well as the "nove" way.

William J. Garner
Yardley, PA 19067



Freq/ MHz	VSWR P1	Loss P1-2	Loss P1-3	Del-L 1-3	Del-P vs. 1-2	VSWR P2	Isol. P2-3	VSWR P3
200.00	1.78	3.55	3.61	.05	−.38	1.21	9.71	1.22
300.00	1.74	3.47	3.51	.04	−.94	1.13	12.15	1.13
400.00	1.59	3.39	3.44	.05	−1.23	1.12	14.77	1.13
500.00	1.42	3.33	3.39	.06	−1.44	1.15	18.44	1.15
750.00	1.04	3.26	3.33	.07	−2.55	1.03	30.97	1.03
1000.00	1.38	3.56	3.63	.07	−3.50	1.25	22.59	1.22
1250.00	1.40	3.56	3.62	.06	−4.19	1.37	27.40	1.32
1500.00	1.06	3.40	3.46	.06	−4.75	1.19	22.72	1.14
1750.00	1.48	3.71	3.78	.07	−5.40	1.45	21.83	1.36
2000.00	1.68	3.88	3.97	.09	−5.73	1.56	25.92	1.46
2250.00	1.20	3.66	3.83	.17	−5.81	1.35	21.04	1.30
2500.00	1.37	3.83	4.11	.28	−5.38	1.64	18.54	1.50
2750.00	2.17	4.22	5.51	1.29	−4.26	3.03	14.08	3.28
3000.00	3.47	5.49	6.51	1.02	−20.50	3.04	9.43	3.79
3250.00	3.09	5.71	5.46	−.24	−15.82	2.15	15.07	2.39
3500.00	1.13	4.29	4.12	−.16	−13.56	1.39	32.92	1.61
3750.00	2.27	5.04	4.90	−.14	−13.45	1.82	19.03	1.83
4000.00	1.62	4.88	4.67	−.21	−13.43	1.60	26.85	1.79

Microstrip layout and measured parameters of the Cohn power divider.

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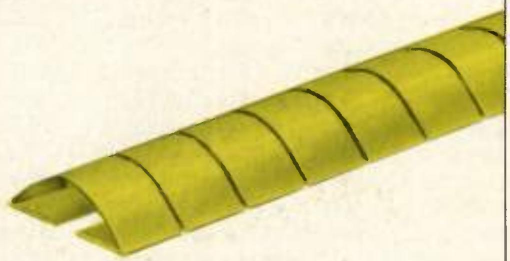
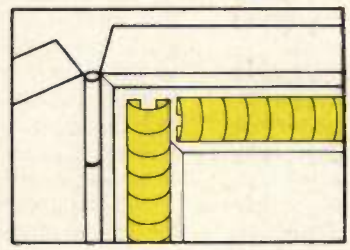


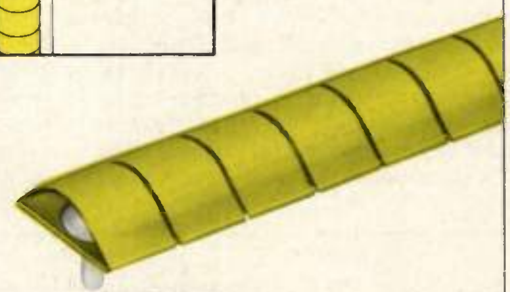
These new symmetrical slotted shielding strips of beryllium copper permit continuous spring contact throughout their length, providing the perfect answer for a variety of shielding requirements.

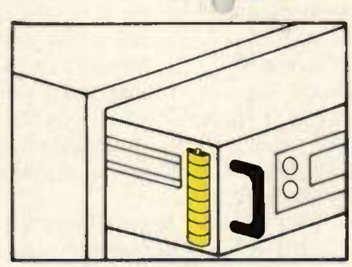
Three models are available: basic, rivet-mount and double-faced adhesive-mount designs. The basic design consists of low-compression, adhesive-mounted strips. A generous radius profile provides for the greatest incident engagement angle with the lowest force. As with all Sticky Fingers® shielding strips, the self-adhesive tape makes mounting easy and secure.



Adhesive-mounted strips for typical electronic enclosures.

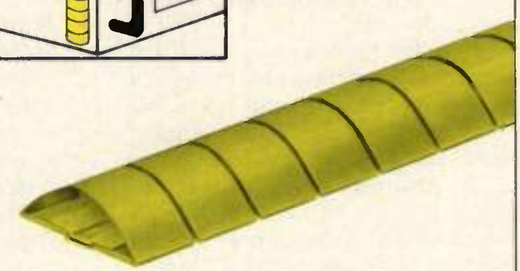
The rivet-mount design incorporates the addition of an integral track, pierced for mounting with nylon push rivets. This configuration allows bi-directional engagement, and is specially designed for slide applications, PC board connections, etc.





Strips with integral mounting track* and nylon push rivets, for sliding drawer applications.

The third design also incorporates an integral track-mount design, but employs a double-faced adhesive tape instead of push rivets. This provides for fast, easy field replacement in military applications, especially where high frequencies do not permit the use of mounting holes.



Strips with integral mounting track* and self-adhesive strip for fast field replacement.

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Australia Completes First Satellite System

Aussat, the Australian national satellite communications system, provides a range of domestic services to the entire continent, its offshore islands, and Papua New Guinea. The service includes television broadcast, television relays between major cities, digital data transmission for telecommunications and business, voice applications, centralized air traffic control, and maritime radio coverage.

According to Hughes Aircraft Co., the satellite builder, Aussat uses two telescoping cylindrical solar panels and a folding antenna for compactness during launch. After the satellite nears its orbital position, the antenna erects and the outer solar panel deploys, exposing the inner solar array. Aussat's dual polarized, three-reflector antenna system provides seven transmit and three receive beams.

The satellite carries 15 channels, each 45 MHz wide. Four use 30 W travelling wave tube amplifiers (TWTAs) to provide radio and television services to Australia's remote areas; the remaining 11 channels operate with 12 W TWTAs. It is possible to connect the communications channels individually to the transmit beams by ground command. This provides traffic assignment flexibility. The electrical power system uses K7 solar cells which provide 1054 W at beginning of life. Two nickel-cadmium batteries provide full power when the spacecraft passes through earth's shadow. The satellites have a mission life of 7 years, operating at the 14/12 GHz Ku band.

The first and second satellites are located just north of Papua New Guinea at 160 degrees E and 164 degrees E longitude. The third satellite is located at 156 degrees longitude and was launched on September 15, 1987. The first two were launched in the system by the U.S. Space Shuttle in August and November 1985.

The third satellite, Aussat 3, is the first Hughes built satellite launched on the Ariane rocket. This satellite is similar to the first two with one exception. It has two horn antennas that provide communication services to New Zealand and other south pacific islands. The master control for the Aussat system is in Sydney, and backup control equipment is in Perth. Monitoring equipment has been installed at earth stations in Sydney, Perth, Brisbane and Adelaide.

With the advent of the Aussat satellites, the Australian continent is covered by a single communication system. 650,000 people living in remote areas are now able to receive television and radio broadcasts for the first time with antennas as small as 4 feet in diameter.

Aussat 3, Australia's third communication satellite built by Hughes. Photo courtesy of Hughes Aircraft Company.

Applications for Superconductors Get Closer

The National Bureau of Standards (NBS), Boulder, Colo., in collaboration with Westinghouse Electric Corp. has devised a method for making improved, lower-resistance electrical contacts on the new high critical temperature ceramic superconductors.

High contact resistance has been a major obstacle to commercial applications of the new superconductors. The new method reduces the contact resistivity several thousand times below that previously achieved with conventional contacts.

Resistance at electrical contacts causes heating in any device, but it is particularly fatal in superconductors. The new contact method has surface resistivities of less than 10 micro-ohm-cm^2, using bulk samples of yttrium-barium-copper-oxide ceramic superconductor ($YBa_2Cu_3O_7$). This level of performance for the "super contacts" was achieved while operating the superconductor at the relatively high temperature of liquid nitrogen-77 K.

A number of contacts have been made using the new method and have been found to be consistently reproducible. Systematic tests conducted on contacts exposed to dry air over a 3-month period showed consistently low resistivity and little degradation with repeated cooling to 77 K and warming to room temperature.

Japan Leads the Way in Superconductor R & D

Despite significant U.S. breakthroughs in superconductivity research, the Japanese appear to be leading the world in developing and implementing a national strategy for high temperature superconductivity research and development. These issues and a range of others are being explored in a multiclient program offered by Battelle. The program, "Superconductivity I: A Technology and Business Assessment," is the first in a series of superconductivity related programs.

"Commercial application of high temperature superconductivity appears to be possible during the next five to ten years," said Battelle's Donald C. Slivka. The initial study is uncovering a wealth of information about the Japanese efforts, he added. Major Japanese firms are investing significant resources in the development of high temperature superconducting material, forms, and devices. For example, Sumitomo Electric Industries, Ltd. has about 65 researchers working with wire and thin film, and 700 patent applications have been filed by another 200 staff members dedicated to superconductivity work.

For more information about Battelle's program, contact Donald Slivka at (614) 424-4090.

NASA Studies Mars

The National Aeronautics and Space Administration has announced Martin Marietta Space Systems of Denver, Colorado as one of two companies contracted by its Jet Propulsion Laboratory to study mobility and surface rendezvous techniques which will allow a robotic rover to navigate and traverse rugged terrain on Mars. In addition, NASA Johnson Space Center awarded a contract to Martin Marietta to study aerobraking to place a spacecraft into Mars orbit, enter its atmosphere, and descend to a landing on

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the planet. The contracts call for soil samples to be returned to earth before the year 2000.

NASA envisions placing spacecraft in orbit around Mars and separating a lander vehicle for descent to the surface. The lander would carry a smaller rover vehicle designed to move over the Martian landscape for about one earth year, selecting surface and subsurface rock and soil samples. The rover would then return to the lander, transferring the collected samples to a small rocket for rendezvous with the original orbiter and subsequent transfer to the U.S. Space Station. After preliminary scientific evaluation aboard the station, the specimens would be returned to earth.

Standard Oil Ceramics Division Wins Industrial Research Award

The Electronics Ceramics Division of Standard Oil Engineered Materials Company received *Research and Development's* Industrial Research (I-R 100) award for producing a new planar diffusion source for n-type arsenic doping in a joint venture with Pennsylvania State University.

The planar diffusion source provides a high degree of doping uniformity and makes it easy to maintain shallow junction depths. The PDS® AS-1000L arsenic source is the latest in the company's PDS line, joining multi-temperature phosphorus and boron nitride dopant materials.

Surface concentrations of 1 x 10^{20} arsenic atoms/cm^3 and junction depths of 0.3 um are easily achieved. Since close control is possible and there is no shadowing effect, the PDS arsenic source is ideal for difficult applications, such as trench capacitors or other structures with high aspect ratios.

NBS Expands NVLAP

The electromagnetic laboratory accreditation program, managed by the National Bureau of Standards (NBS) under the National Voluntary Laboratory Accreditation Program (NVLAP), has been expanded to meet requests from participating accredited laboratories, the producers of electronic equipment, and the U.S. Naval Air Systems Command. Test methods have been added by NBS to help laboratories improve the quality of the testing services on products that must meet FCC approval.

The expanded program includes test methods for RF devices, including receivers (FCC Part 15), industrial, scientific, and medical devices (FCC Part 18), and radio transmitters (FCC Part 90). At the request of the Naval Air Systems Command, NBS added a military standard (MIL STD-462) for the measurement of electromagnetic interference characteristics.

Laboratories interested in accreditation for any of the test methods offered under the expanded program or for information on NVLAP should contact Harvey Berger at (301) 975-4016.

AIM Promotes Auto ID Curricula

Automatic Identification Manufacturers, Inc., cosponsored a program with Ohio State University to educate professors and instructors about automatic identification and provide them with a foundation for teaching their students about Auto. ID and its potential to the business world.

Auto ID technology includes the familiar bar codes, RF identification and various other forms. The RF identification technology employs bidirectional radio signals as the encoding medium. It is widely used to provide hands-free access control for vehicle identification in the transport industry and in a host of industrial automation and material handling applications where there is no line of sight between scanner and ID tag. Further information can be obtained by calling AIM at (412) 963-8588.

Wescon Challenger Scholarship Announced

Michael E. Clayton, a senior at Bishop High School, Boise, Idaho, has been selected winner of the $5,000 Wescon Challenger Scholarship for 1987. Begun in 1986, the scholarship is awarded by the Wescon Board of Directors in honor of the seven astronauts who died in the explosion of the Challenger shuttle.

Thomson Components-Mostek Plans Advanced Fab Line

Pasquale Pistorio, President of the combined SGS and Thomson semiconductor companies, announced an initial $12 million capital investment in the Thomson Components-Mostek Corporation manufacturing plant in Carrollton, Texas. With the investment, the existing advanced technology wafer fabrication line will be equipped for early 1988 volume 6-inch production. The line will be initially used for manufacturing ICs with 1.2 micron double- level metal HCMOS technology and will be able to go down to submicron technologies.

CE Opens New Facility

Cincinnati Electronics officially opened its new facility in Mason, Ohio on October 11, 1987 which houses its growing Aerospace Division. About half the 93,000 square feet of floorspace is dedicated to eight laboratories designed for research and development. Plans call for the C³I division to move from its present location to the new campus-style environment in Mason by 1993.

Fluke and Philips Create Global Alliance

The third largest global organization in the electronic test and measurement industry came into being on September 28, 1987 with the signing of a long term agreement creating an alliance between John Fluke Mfg. Co., Inc. (U.S.A.) and Philips T & M (Netherlands). The terms of the agreement were reported in the June issue of *RF Design* (p. 14). In addition to future independent product development programs, the companies will also consider joint ventures to develop products lines in categories in which neither is a significant participant today.

GE Integrates Intersil and RCA Lines

GE Solid State has fully integrated its lines of Intersil and RCA digital signal processing devices and data converters. Engineering and marketing responsibility for the combined lines will be at the Intersil headquarters facility in Cupertino, Calif. Added to the Intersil DSP line are three former RCA products: the CDSP100, a 40-tap, 20 MHz FIR filter; the CDSP110, a 10 MHz LMS adaptive FIR filter; and the CDSP200, a video FIFO. The products will be renamed ISP9400, ISP9410, and ISP9500, respectively. In addition, Intersil assumes responsibility for an RCA data conversion product line that includes the CA3300 Series of 4 to 8-bit flash converters.

rf courses

The George Washington University

Tools for Evaluating Command, Control, and Communication Systems
December 2-4, 1987, Washington, DC

Wideband Communications Systems
December 7-11, 1987, Washington, DC

Modern Digital Signal Processing and Applications
December 14-18, 1987, Washington, DC

SAW Devices and their Signal Processing Applications
February 22-25, 1988, Washington, DC

Electromagnetic Interference and Control
February 22-26, 1988, Washington, DC

Fiber-Optics System Design
February 29-March 2, 1988, Washington, DC

Hazardous RF Electromagnetic Radiation
March 16-18, 1988, Washington, DC

Microwave Radio Systems
March 7-8, 1988, Washington, DC

Frequency-Hopping Signals and Systems
March 21-23, 1988, Washington, DC

Spread-Spectrum Communications Systems
April 4-8, 1988, Washington, DC

Grounding, Bonding, and Shielding
April 7-8, 1988, Washington, DC

Modern Communications and Signal Processing
April 18-22, 1988, Washington, DC

Introduction to Receivers
April 18-19, 1988, Washington, DC

Modern Receiver Design
April 20-22, 1988, Washington, DC

Information: Shirley Forlenzo, Continuing Education Program, George Washington University, Washington, DC 20052; Tel:(800) 424-9773, (202) 994-8530

Besser Associates

Principles of RF Design — Theory and Applications
Dec. 14-16, 1987, Santa Clara, CA

Microwave Circuit Design I: Linear Circuits
Feb. 1-5, 1988, Los Angeles, CA

Microwave Circuit Design II: Non-linear Circuits
Feb 8-12, 1988, Los Angeles, CA

Information: Les Besser, Besser Associates, Inc., 3975 East Bayshore Road, Palo Alto, CA 94303; Tel: (415) 969-3400

UCLA Extension

Submicron Electronic Devices
January 5-7, 1988, Los Angeles, CA

Superconductive Electronics
January 26-28, 1988, Los Angeles, CA

Microwave Circuit Design I
February 1-5, 1988, Los Angeles, CA

Microwave Circuit Design II
February 8-11, 1988, Los Angeles, CA

Information: UCLA Extension, P.O. Box 24901, Los Angeles, CA 90024; Tel:(213) 825-1901; (213) 825-1047; (213)825-3344

Test Systems, Inc.

MIL-STD-1553
Dec 8-9, 1987, Phoenix, AZ
Feb 24-25, 1988, Phoenix, AZ
May 10-11, 1988, Phoenix, AZ

Information: Leroy Earhart, Test Systems, Inc., 217 W. Palmaire, Phoenix, AZ 85021; Tel: (602) 861-1010

Interference Control Technologies, Inc

Grounding and Shielding
January 26-29, 1988, Orlando, FL

Tempest Facilities
January 19-22, 1988, Phoenix, AZ

Intro to EMI/RFI/EMC
Dec 8-10, 1988, Washington, DC
January 23-25, 1988, Atlanta, GA

Information: Penny Caran, Registrar, Interference Control Technologies, Inc., State Route 625, P.O. Box D, Gainsville, VA 22056; Tel:(703) 347-0030

Integrated Computer Systems

Digital Signal Processing
December 8-11, 1987, Boston, MA
January 26-29, 1988, Washington, DC

Image Processing and Machine Vision
January 19-22, 1988, Los Angeles, CA
February 9-12, 1988, Washington, DC

Hands-On Programming in C
December 8-11, 1987, Palo Alto, CA
December 15-18, 1987, Washington, DC

Hands-On Advanced Programming in C
December 8-11, 1987, Los Angeles, CA
January 19-22, 1988, Washington, DC
February 23-26, 1988, San Diego, CA
March 8-11, 1988, Ottawa, Canada
April 5-8, 1988, Washington, DC
April 19-22, 1988, Los Angeles, CA
May 10-13, 1988, Toronto, Canada

Fiber Optic Communication
December 8-11, 1987, Palo Alto, CA
December 15-18, 1987, Washington, DC

Information: Barbara Fischer, Integrated Computer Systems, 5800 Hannum Avenue, P.O. Box 3614, Culver City, CA 90321-3614; Tel:(800) 421-8166, (213) 417-8888

R & B Enterprises

Understanding and Applying MIL-STD-461C
December 14-15, 1987, Philadelphia, PA

Electromagnetic Pulse Workshop
December 11, 1987, Philadelphia, PA

MIL-STD-461C Praxis (Workshop)
December 16-17, 1987, Philadelphia, PA

Information: Greg Gore, Director of Training, R & B Enterprises, 20 Clipper Road, West Conshohocken, PA 19428. Tel: (215) 825-1960

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30th ARFTG
Crowne Plaza Holiday Inn, Dallas, TX
Information: Ken Bradley, M/S 255, Texas Instruments, P.O. Box 660246, Dallas, TX 75266; Tel: (214) 995-6158

January 11-14, 1988
SMART IV
Westin Bonaventure, Los Angeles, CA
Information: EIA, 2001 Eye St., N.W., Washington, DC 20006; Tel: (202) 457-4932

January 20-21, 1988
San Diego Electronics Show
Del Mar Fairgrounds, San Diego, CA
Information: Epic Enterprises, Show Management, 3838 Camino Del Rio North-Suite 164, San Diego, CA 92108; Tel: (619) 284-9268

February 7-9, 1988
ADEE 88
Rivergate Exhibition Center, New Orleans, LA
Information: Show Manager, ADEE West, Cahners Exposition Group, 1350 East Touhy Ave., P.O. Box 5060, Des Plaines, IL 60017-5060. Tel: (312) 299-9311

February 10-12, 1988
RF Technology Expo '88
Disneyland Hotel, Anaheim, CA
Information: Linda Fortunato, Cardiff Publishing Company, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111; Tel: (303) 220-0600; (800) 525-9154

February 23-25, 1988
NEPCON West '88
Anaheim Convention Center, Anaheim, CA
Information: Jerry Carter, Cahners Exposition Group, 1350 East Touhy Ave., P.O. Box 5060, Des Plaines, IL 60017-5060; Tel: (312) 299-9311

March 8-10, 1988
Southcon '88
Orange County Convention Center, Orlando, FL
Information: Electronic Conventions Management, 8110 Airport Boulevard, Los Angeles, CA; Tel: (213) 772-2965

May 10-12, 1988
Electro '88
Bayside Exposition Center, Boston World Trade Center, Boston, MA
Information: Electronic Conventions Management, 8110 Airport Boulevard, Los Angeles, CA; Tel: (213) 772-2965

May 25-27, 1988
1988 IEEE MTT-S International Microwave Symposium
Javitts Auditorium, New York City, NY
Information: Charles Buntschuh, Narda Microwave Corp., 435 Moreland Road, Hauppauge, NY 11788; Tel: (516) 231-1700

June 1-3, 1988
42nd Annual Frequency Control Symposium
Stouffer Harborplace Hotel, Baltimore, MD
Information: Raymond L. Filler, Frequency Control and Timing Branch, Department of the Army, Electronics Technology and Devices Laboratory, Fort Monmouth, NJ 07703-5000

New ADCs for RF Signal Processing

By Neil Albaugh
Burr-Brown Corporation

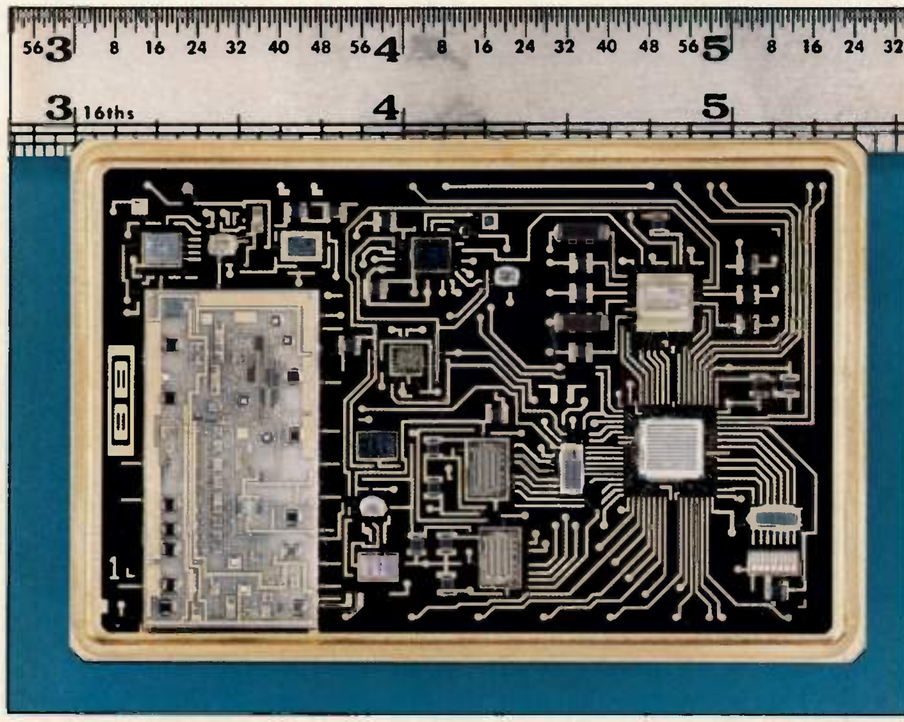
Digital signal processing techniques offer the potential for unprecedented performance in radar, ECM, and communications receivers. Powerful signal processing functions, which have not been readily achievable by analog means, are now available to the RF designer. This trend is accelerating as high speed digital devices continue rapid development.

Digital RF processing requires two key components to interface with the analog RF world: an analog-to-digital converter (ADC) and a digital-to-analog converter (DAC). High dynamic range system requirements place severe performance demands on these devices, requiring both high resolution and low distortion. Recent advances in technology have dramatically improved the performance of ADCs and DACs, particularly with regard to speed and resolution.

Early designs utilized flash ADCs but the quantizing noise of these low resolution converters severely limited performance. Currently, high performance systems use 12 bit converters to achieve low noise and, consequently, to improve signal to noise ratio (SNR).

High-speed 12 bit converters with good performance are currently available on the market, but they have much room for improvement in terms of size and power consumption. Flash converters are not available at the 12 bit level due to the unmanageable circuit complexity that is required by this approach. Current high-performance ADCs use a sub-ranging architecture (Figure 1) to achieve high resolution at a high conversion speed.

Burr-Brown's new 12 bit, 10 megahertz conversion rate analog-to-digital converter, the ADC603, is a significant development for RF signal processing applications. It offers small size, hermetic package, low power consumption and wide temperature range, while matching or exceeding the performance of an older generation of large modular ADCs. The ADC603 uses hybrid construction techniques with a very high level of integration to pack an entire sub-system including buffer, sample/hold amplifier, reference, and a sub-ranging ADC into a 46-pin ceramic and metal DIP package. Power dissipation is only 4.5 watts.



Burr-Brown's ADC603 offers speed and resolution in a small hybrid package.

High-Speed ADC Design

To help understand the sub-ranging ADC technique, a detailed block diagram of a typical sub-ranging converter, the modular ADC600, is shown in Figure 2. An input buffer drives a sample/hold which captures an "instantaneous" snapshot of the input signal amplitude. The voltage held by the S/H is then converted to digital by a 7 bit "MSB" flash ADC. The coarsely digitized signal is then converted back to an analog voltage by a DAC. While this DAC has only 7 bit resolution, its accuracy and linearity approach 14 bit levels. The DAC analog output then drives a differential amplifier where it is subtracted from the original S/H output signal. In this way, an error signal is generated which corresponds to a "remainder" from the first coarse digital conversion.

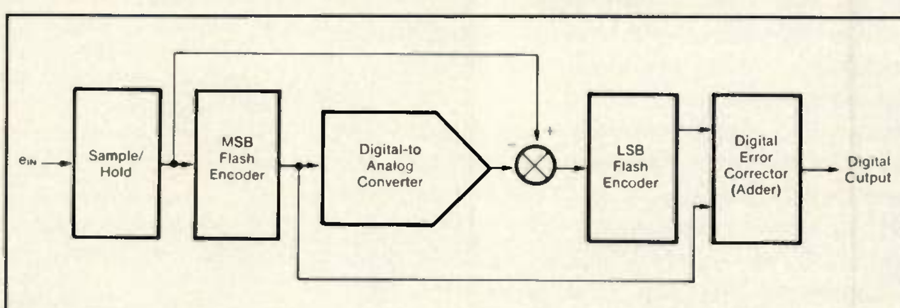


Figure 1. Sub-ranging analog-to-digital converter block diagram.

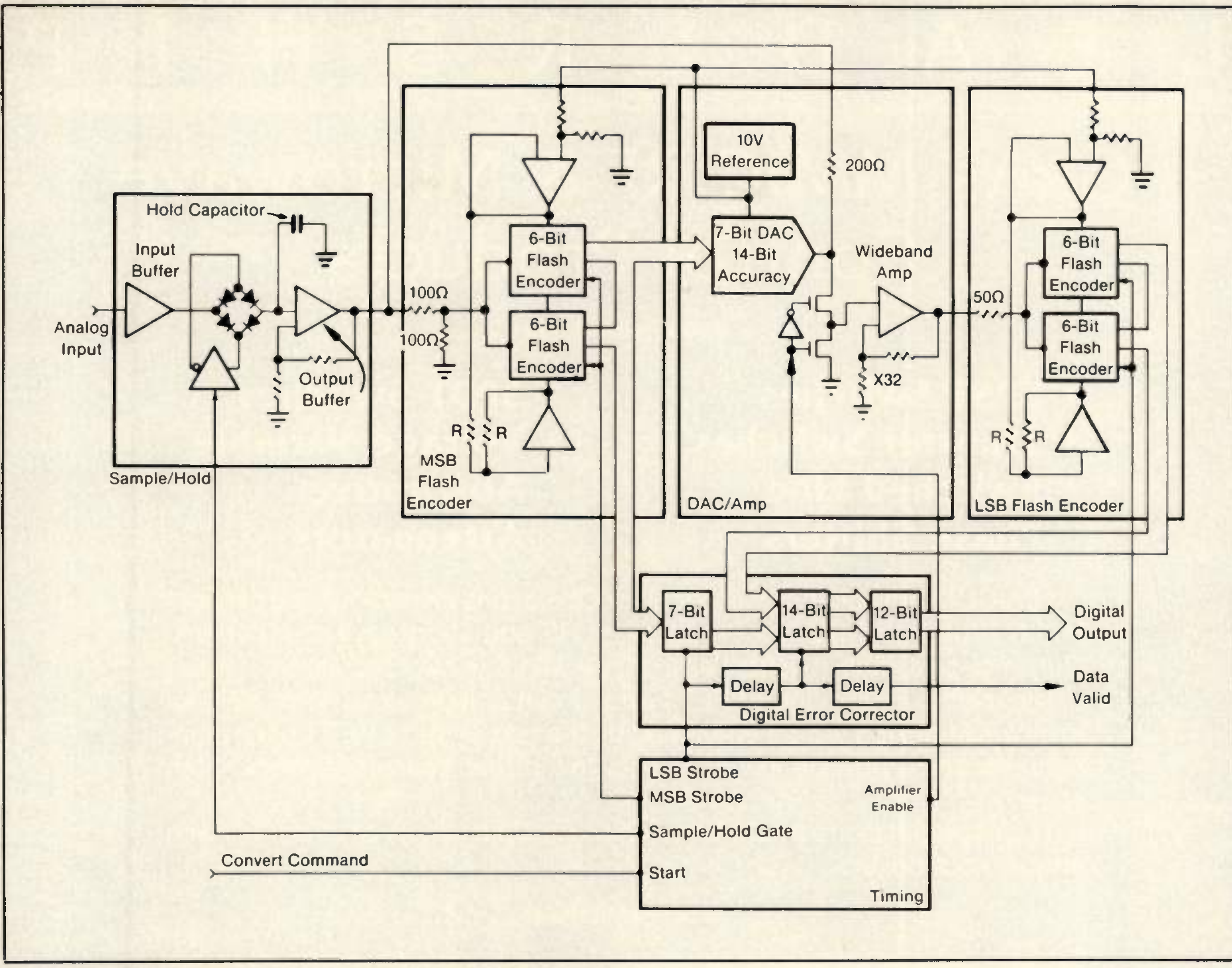


Figure 2. Block diagram of the 12-bit 10 MHz ADC600.

This analog remainder is then amplified and converted to digital by a second 7 bit "LSB" flash ADC. The outputs from both the "MSB" and "LSB" flash converters are combined in a digital error correction circuit which yields a 12 bit digital output and also senses an over-range condition. This prevents a data output "roll over" from occurring under overload conditions.

The 7 bit flash ADC circuit function in the ADC600 and in the new ADC603 is achieved by a parallel connection of two 6 bit flash ADCs as shown in Figure 3. To minimize power consumption in the new converter, one "7 bit" flash converter is multiplexed to allow both MSB and LSB conversions and new proprietary approaches to sub-ranging have been utilized. Besides 12 bit resolution, the ADC603's high-performance includes excellent linearity and exceptionally low aperture uncertainty (aperture jitter).

Linearity, the key parameter in achieving low distortion, has been difficult to achieve in high speed converters. Two

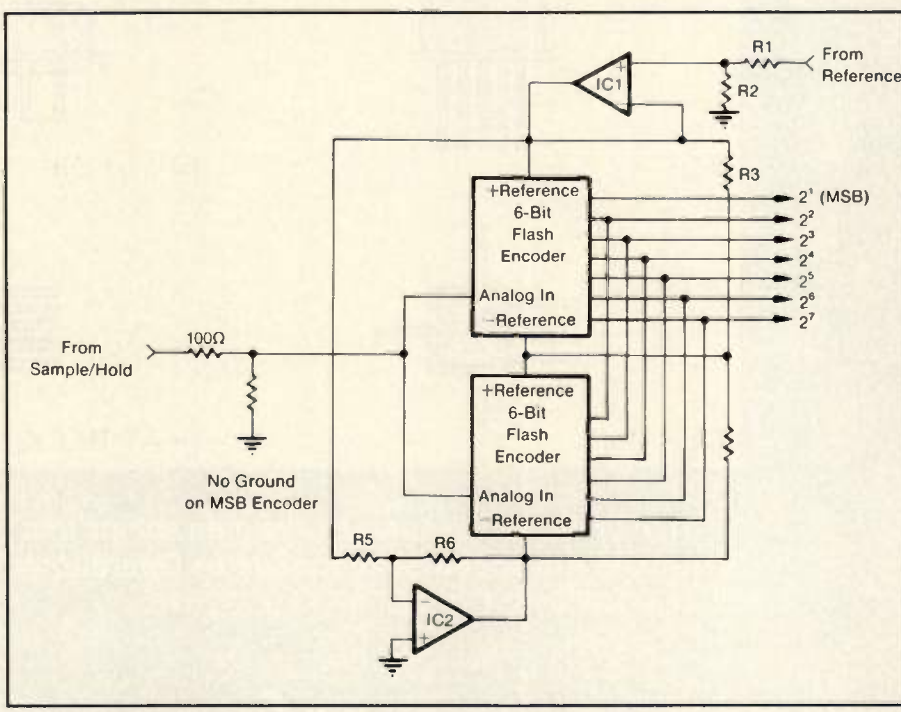


Figure 3. Seven-bit flash encoder.

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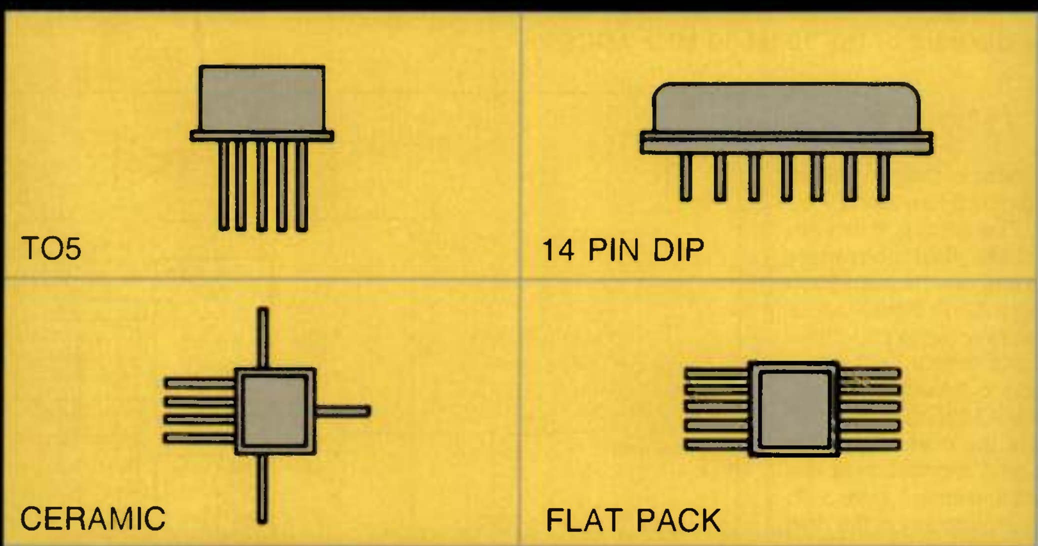

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linearity specifications are normally encountered by data converter users. First, differential non-linearity (DNL) refers to the error of each individual bit transition from its ideal value; DNL contributes excess quantizing noise. Second, integral non-linearity (INL) refers to the departure from a true linear transfer function; INL contributes harmonic and intermodulation distortion.

While the linearity of an ADC is important, its sample/hold amplifier is of equal, if not greater, importance. Low distortion is vital in a high performance sample/hold, as it will add its non-linearity to the basic non-linearity of the ADC.

Analog-to-digital converters used for RF signal processing applications should have low noise (a function of resolution and DNL) and low harmonic and intermodulation distortion (a function of INL). Modern ADCs exhibit these characteristics (1). An FFT spectral plot of a 12 bit ADC driven by a full-scale (envelope) two-tone test signal is shown in Figure 4. As the sampling rate was 5 MHz, the FFT displays a signal spectrum of 0 to 2.5 MHz, or 1/2 the sampling frequency F_s. Higher frequency distortion products are displayed but they are aliased to a frequency within this FFT display.

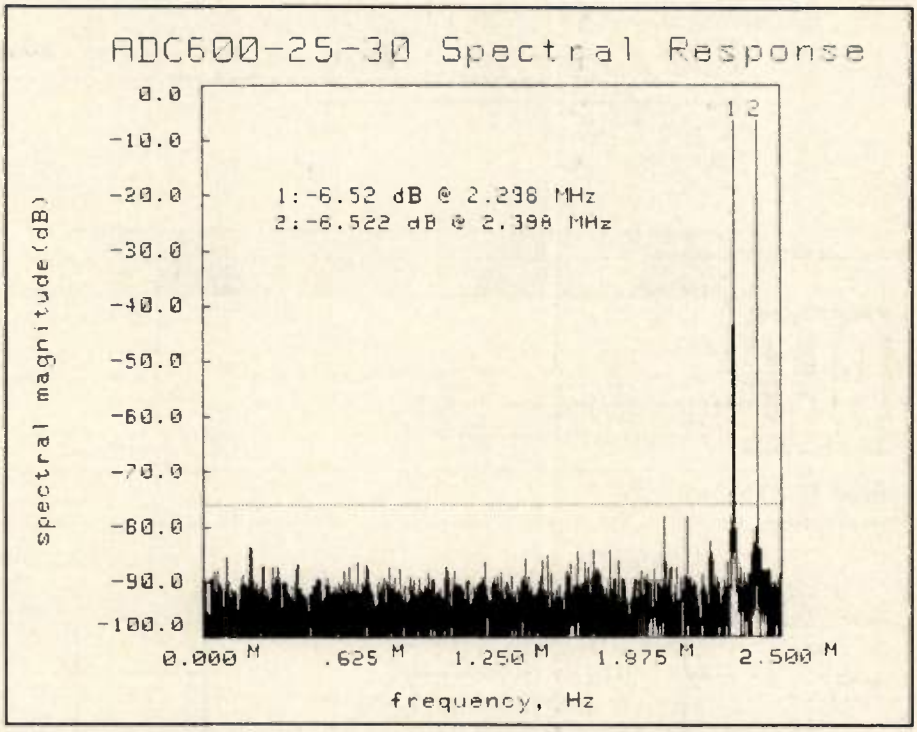


Figure 4. Two-tone intermodulation distortion performance of an RF signal processing ADC.

Sampling Theory and Aliasing

"Aliasing" may sound mysterious to an RF engineer but a look at basic sampling theory reveals familiar principles disguised by different terminology. The operation of a sampling ADC (Figure 5) can be understood by referring to the spectral plots of Figure 6. A narrow sampling impulse F_s, generates a series of harmonics which, when convolved with an input signal F, results in a series of double-sideband suppressed carrier amplitude modulated harmonics $F_s \pm F$.

This is similar to the familiar harmonic mixer (Figure 7). At a signal frequency F higher than $F_s/2$, adjacent upper sidebands (USB) and lower sidebands (LSB) will overlap. This is the mysterious "aliasing error"; $F = F_s/2$ is the Nyquist limit. Although unambiguous conversion re-

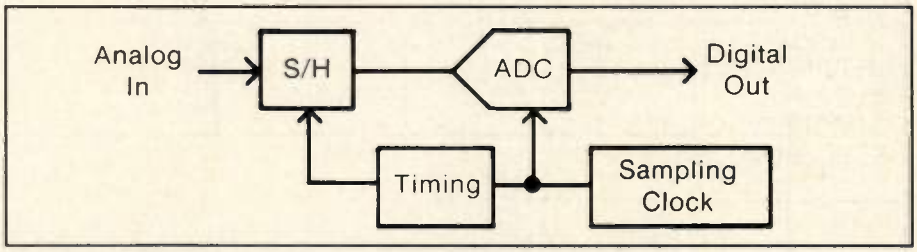


Figure 5. Sampling analog-to-digital converter (ADC).

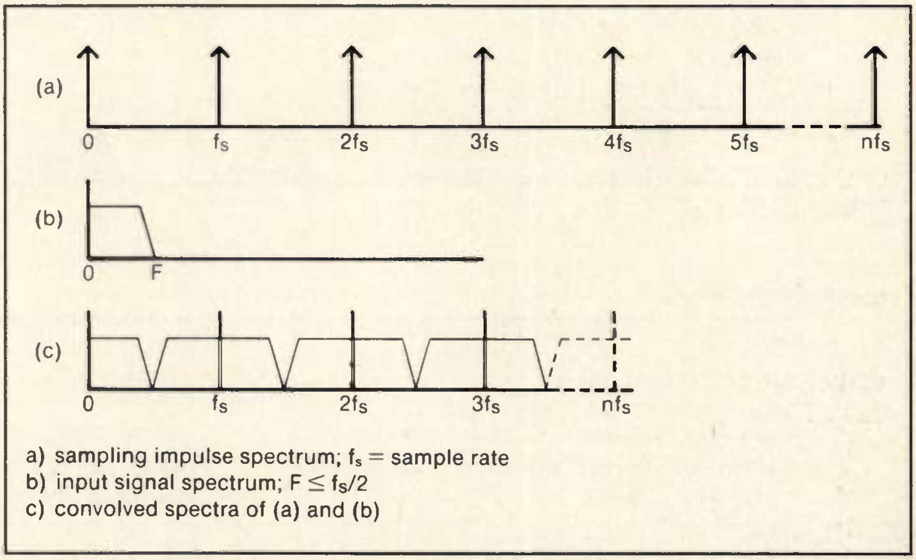


Figure 6. Sampling theory.

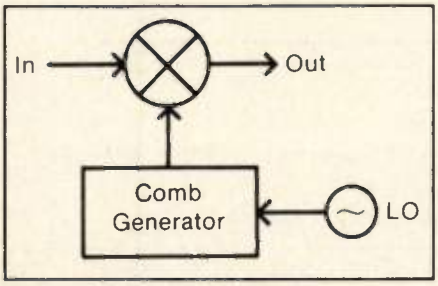
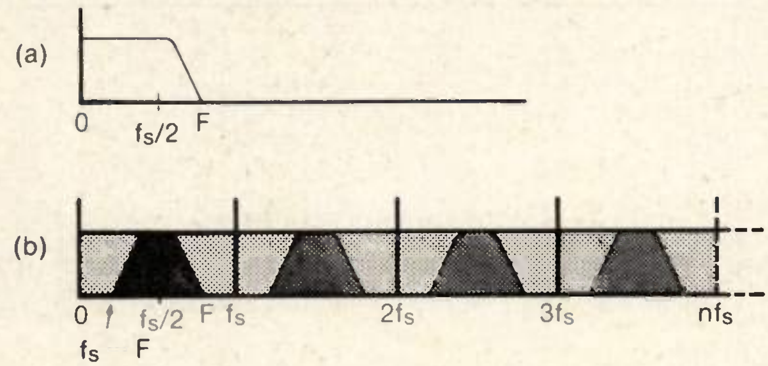


Figure 7. Harmonic mixer.



a) input signal spectrum; $F > f_s/2$
b) spectral overlap: "aliasing"

Figure 8. Aliasing error.

quires a sample rate of at least twice a baseband input signal frequency to prevent aliasing, a much higher frequency input signal can be digitized under the proper conditions (2).

Harmonic sampling can be accomplished with an ADC by using an approach such as Figure 9. The important condition for avoiding aliasing is that the input signal bandwidth be less than $F_s/2$. The Nyquist criteria is not violated provided no USB/LSB overlap occurs. As seen in Figure 10, a 30 MHz to 35 MHz input signal is down-converted by the ADC to a 0 to 5 MHz baseband output signal. Lower-sideband conversion is possible but the output signal will have an inverted spectrum. The digital sampling technique shown in Figure 9 can be recognized as

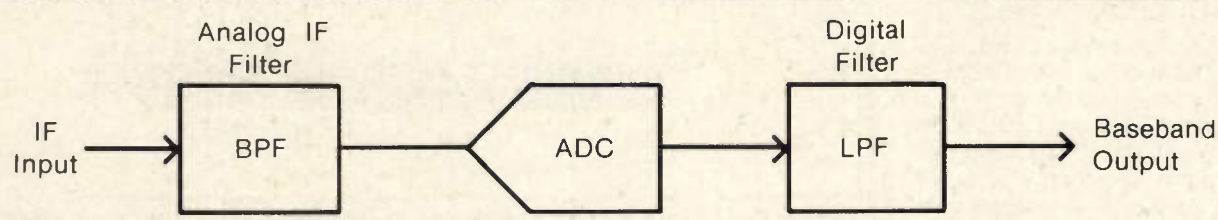
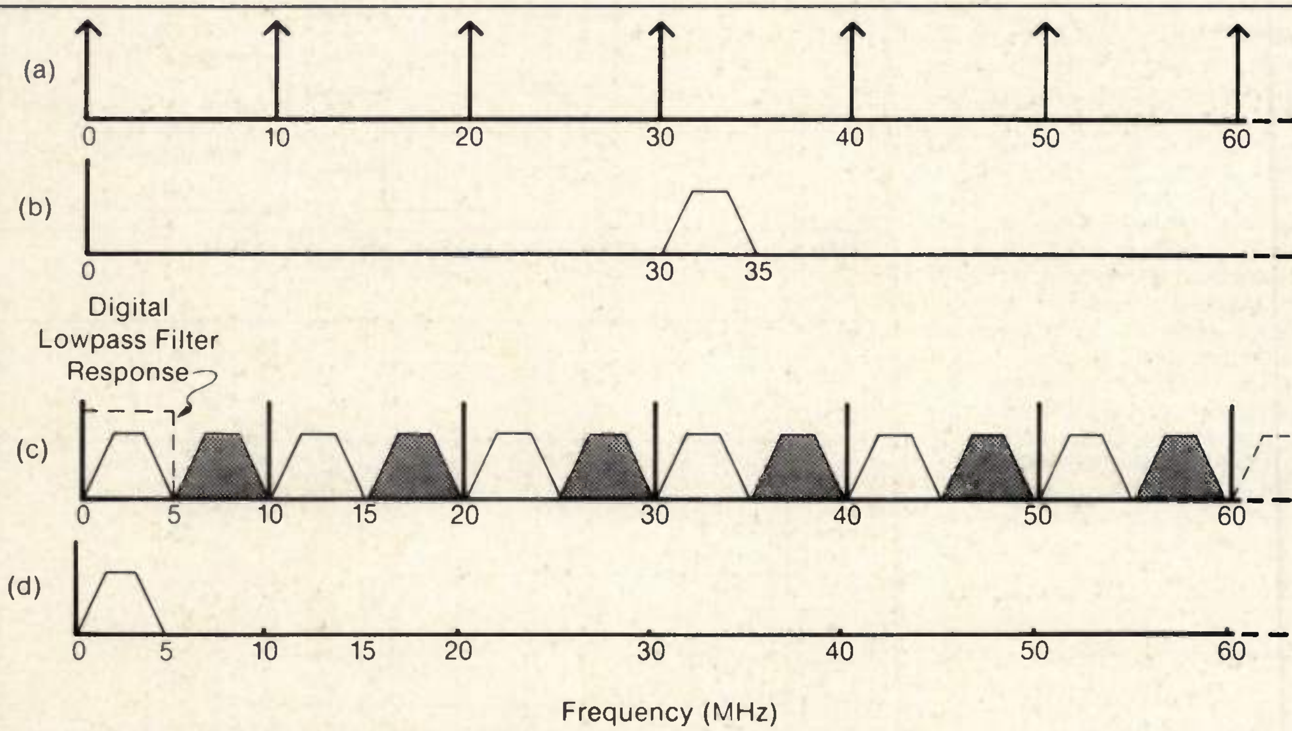


Figure 9. IF frequency conversion by harmonic sampling.



a) sampling impulse spectrum; $f_s = 10MHz$
b) bandwidth limited signal; $F = 30$ to $35MHz$
c) convolved spectra of (a) and (b); shaded sideband has inverted spectrum
d) digitally filtered baseband output

Figure 10. Harmonic sampling.

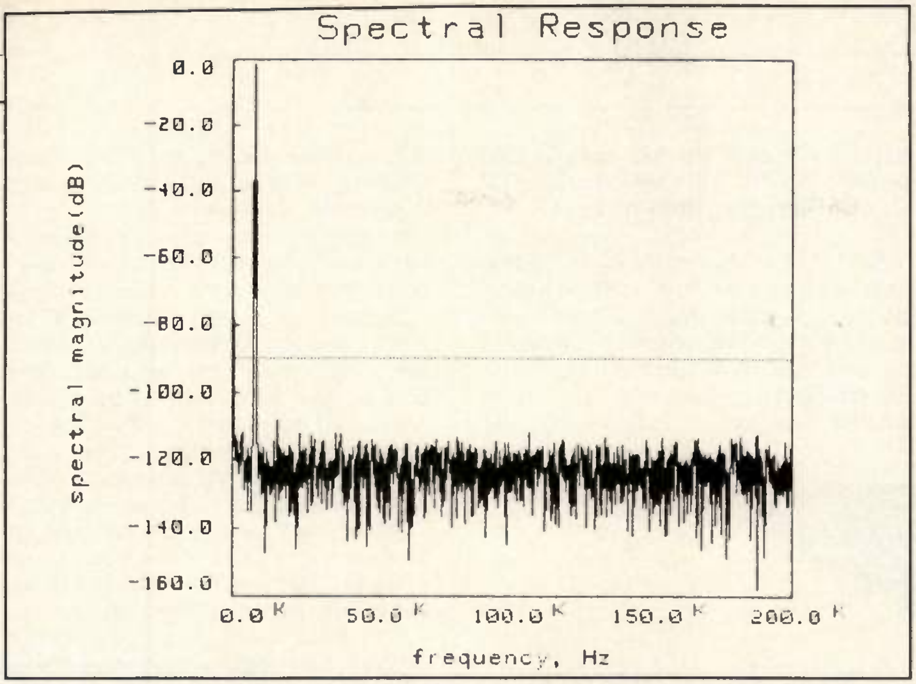


Figure 11. A new high dynamic range 16-bit converter, the ADC701.

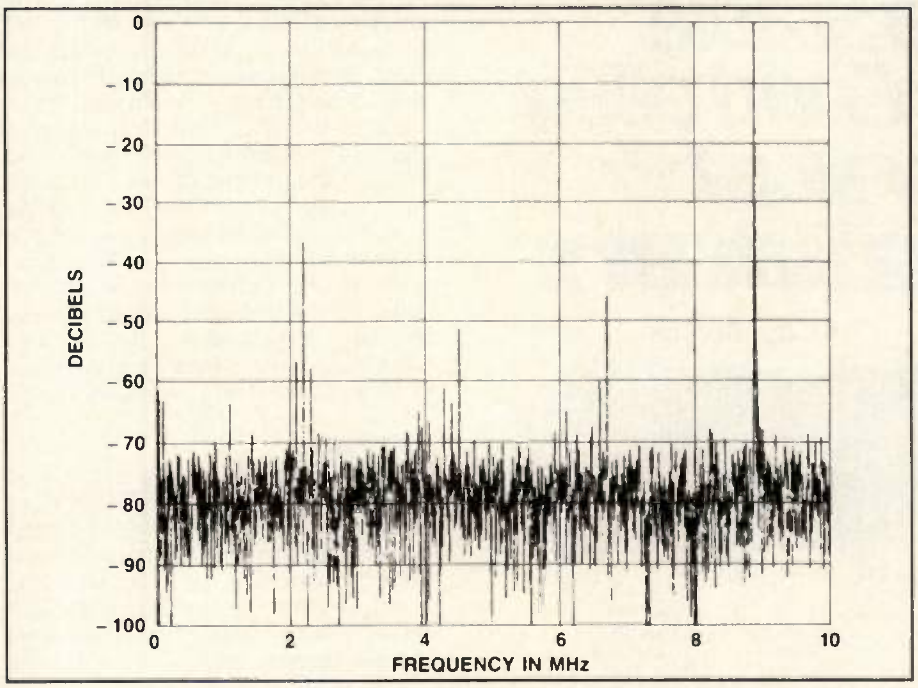
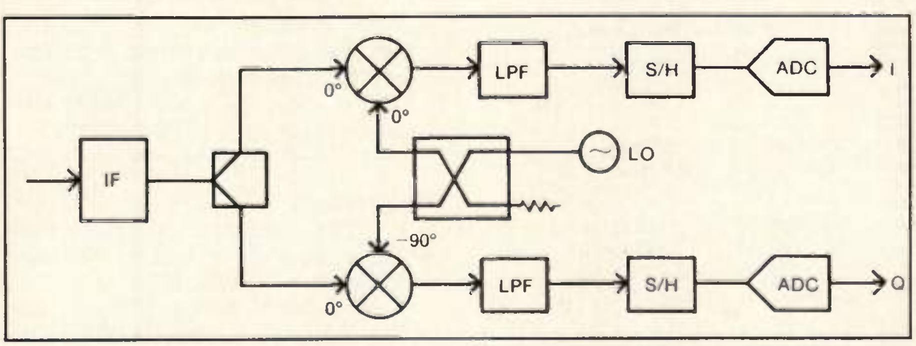


Figure 12. Low-resolution flash converters compromise dynamic range.



Example of RF signal processing application: Vector demodulation of in-phase and quadrature signals.

a variation of the superhetrodyne receiver.

Since the IF can extend down to DC, a homodyne (zero frequency IF) receiver is also easily implemented with an ADC. These receivers, as well as Doppler radar systems and some sonar systems can generate high dynamic range signals that are beyond the capabilities of most ADCs. Very high resolution 16 bit converters (sometimes also employing oversampling) are used to lower the quantizing noise floor. The new 450 kHz sample rate ADC701 and SHC702 show distortion products of better than −100 dBc! (See Figure 11.)

Selection of converters for RF signal processing is not without pitfalls, however. Some high-speed ADCs are not well suited to harmonic sampling. Obviously, the S/H must have adequate analog input bandwidth to pass the input signal without distortion. In addition, very low S/H aperture uncertainty is required to preserve a good S/N ratio. Aperture uncertainty, sometimes referred to as aperture jitter, is analogous to phase noise modulation of an LO signal. This causes noise modulation of the converted signal.

Sampling at a high harmonic of the ADC sample frequency places severe demands on its S/H amplifier. Burr-Brown's new SHC601BH is particularly well-suited for harmonic sampling applications as it features a −3 dB analog input bandwidth of 120 MHz and an aperture uncertainty of only 1 ps RMS.

An excellent sample/hold can even improve a poor ADC under certain conditions. For example, the noise and distortion of flash ADCs suffer at very high conversion rates due to unequal propagation delays in the hundreds of comparators used to sense the input signal amplitude. With a large amplitude high frequency input signal, slew rate induced errors caused by these differential delays become serious, generating missing codes and causing linearity problems, which translate into poor noise and distortion (Figure 12). By adding a fast sample/hold ahead of the flash ADC, a sample of the input voltage can be held fixed for the flash ADC, thus eliminating propagation delay errors due to slew rate. At high frequencies improvements of 10 to 15 dB are achievable by adding a good sample/hold to a flash ADC. This is a powerful and often unrecognized technique, but it has been well described in literature (3).

Conclusion

Digital signal processing will expand the choices available to the RF engineer. As faster ADCs, DACs, and S/Hs are developed, the digital RF signal processor will move closer to the antenna. In the future, we will see both analog and digital RF processing assume their proper roles in RF system designs. rf

About the Author

Neil Albaugh is Product Line Manager in the Advanced Products Division of Burr-Brown Corporation, P.O. Box 11400, Tucson, AZ 85734. He can be reached at (602) 746-7213.

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3. J. Neal and J. Surber, in *Electronic Design*, May 3, 1984, pp. 381-388.

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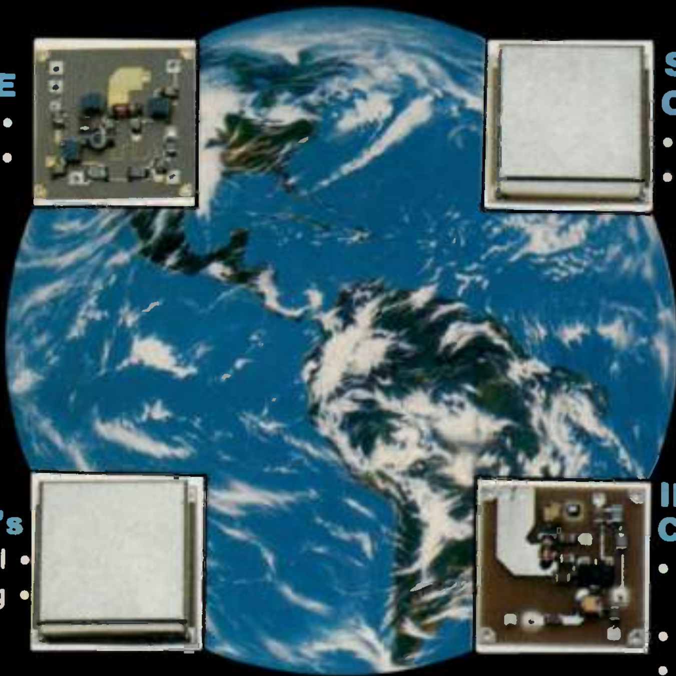

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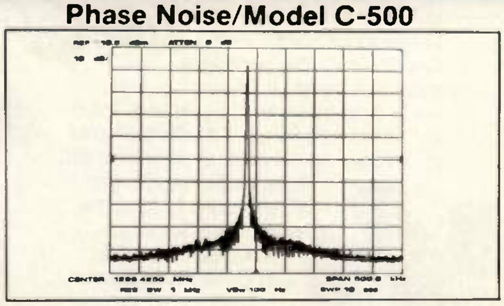

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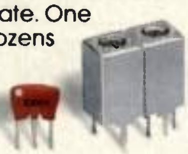

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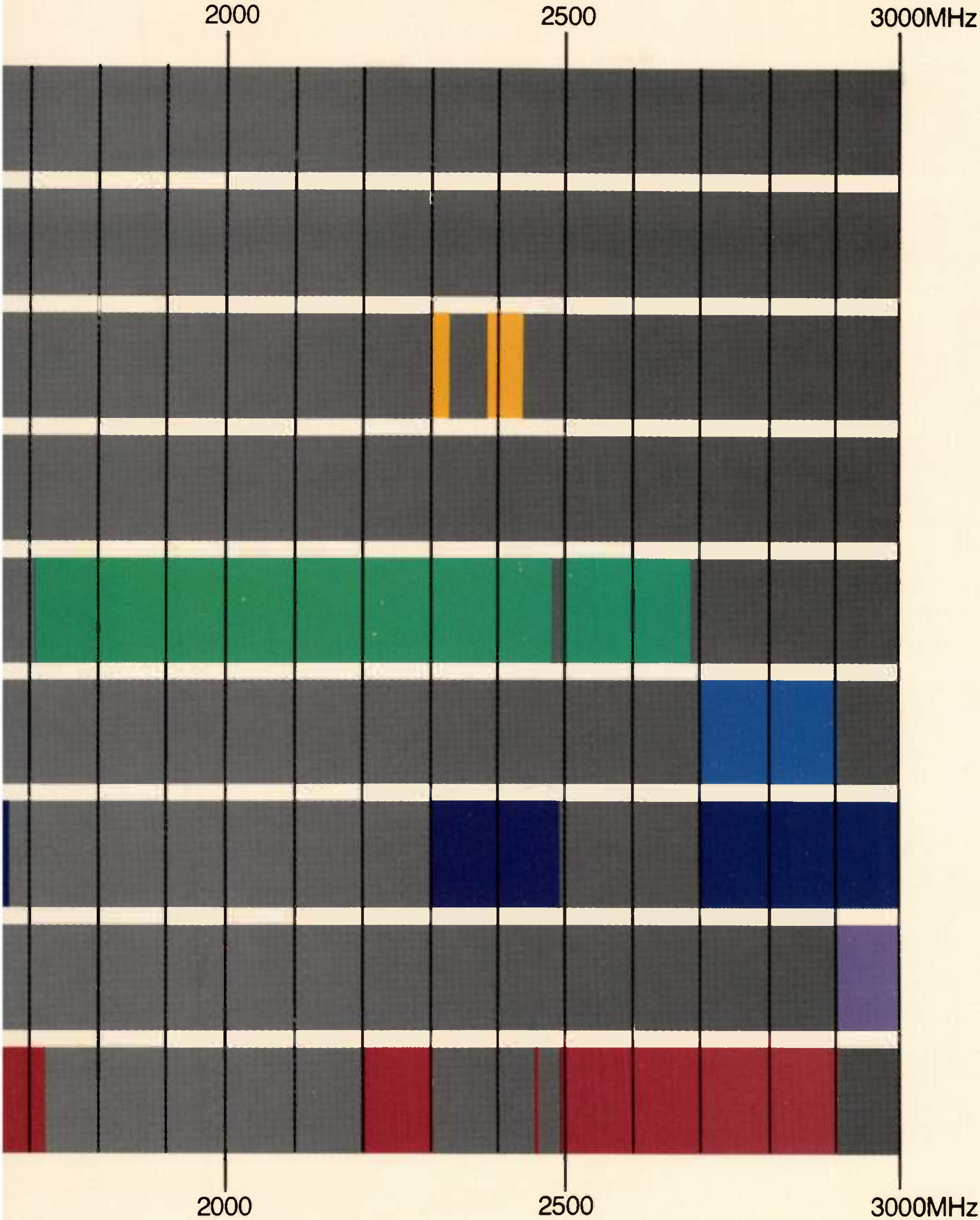
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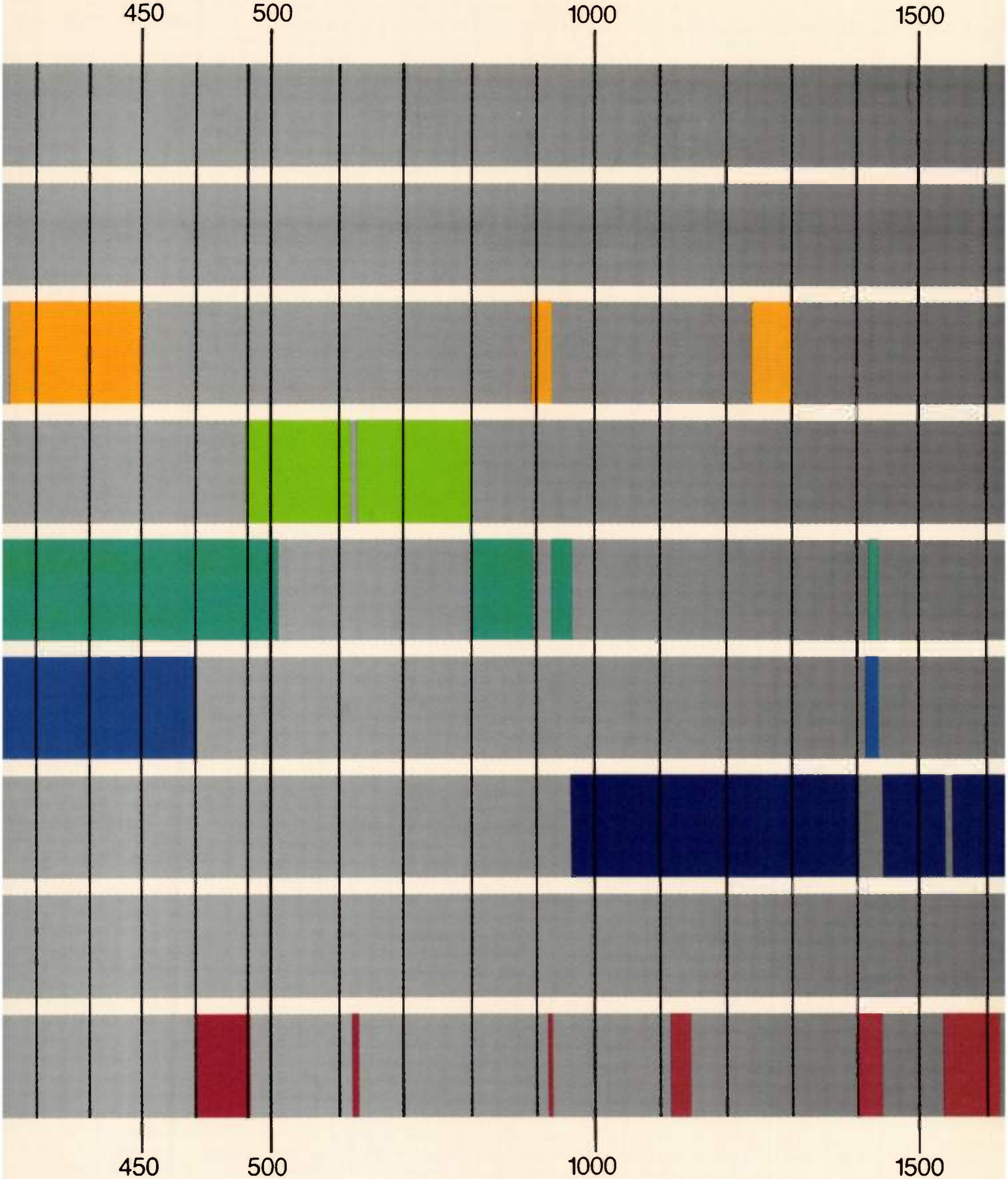

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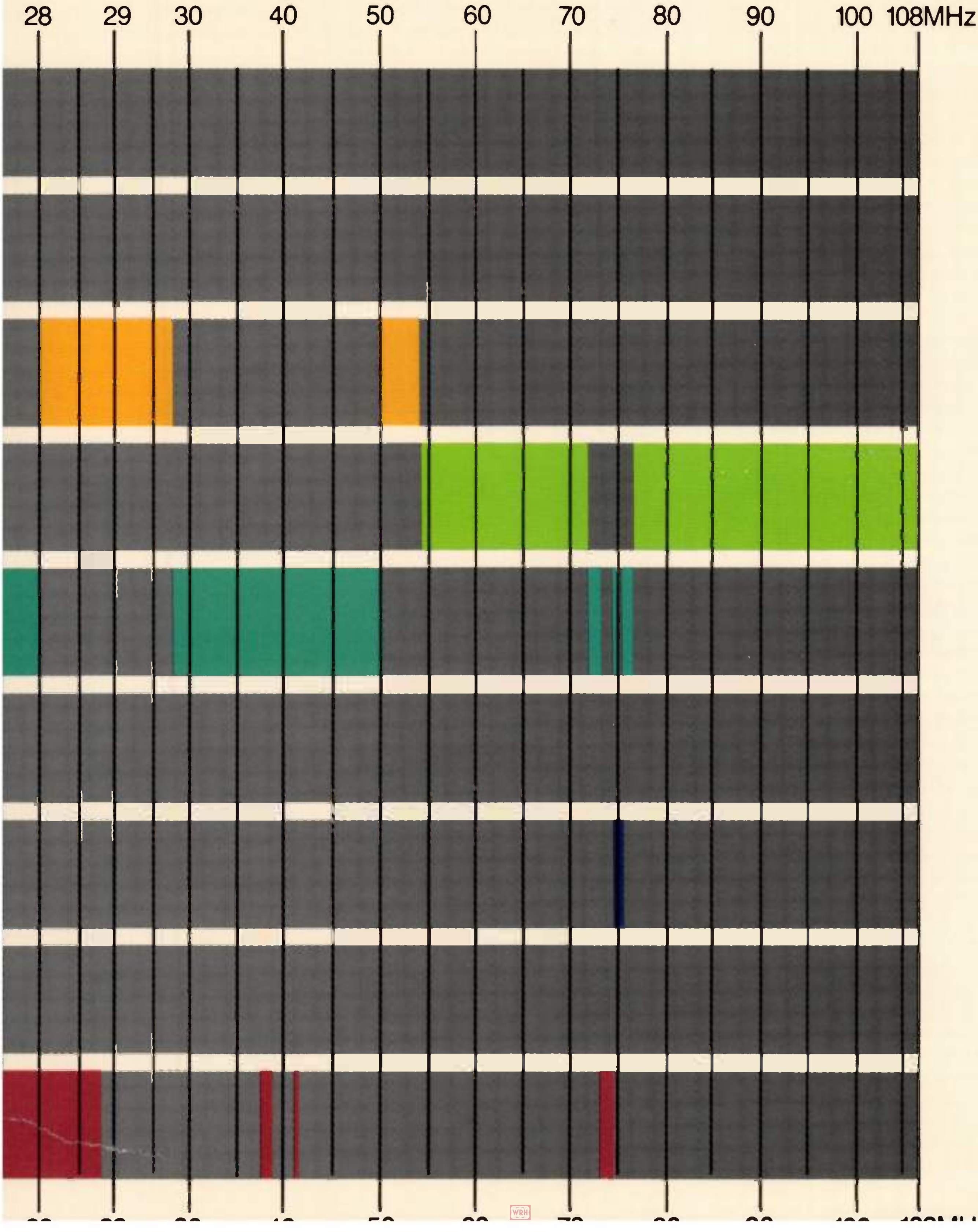

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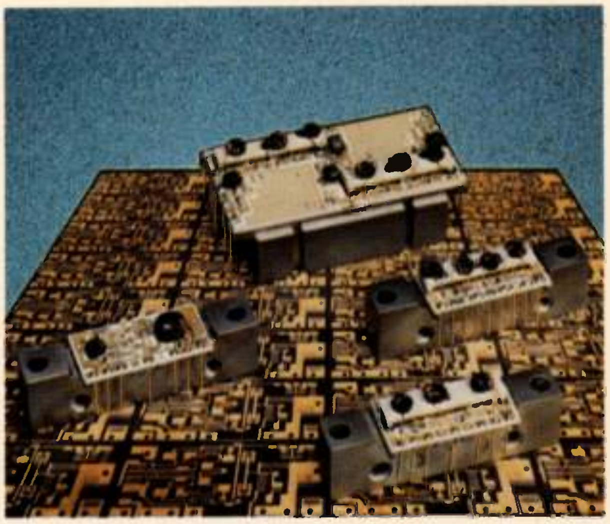

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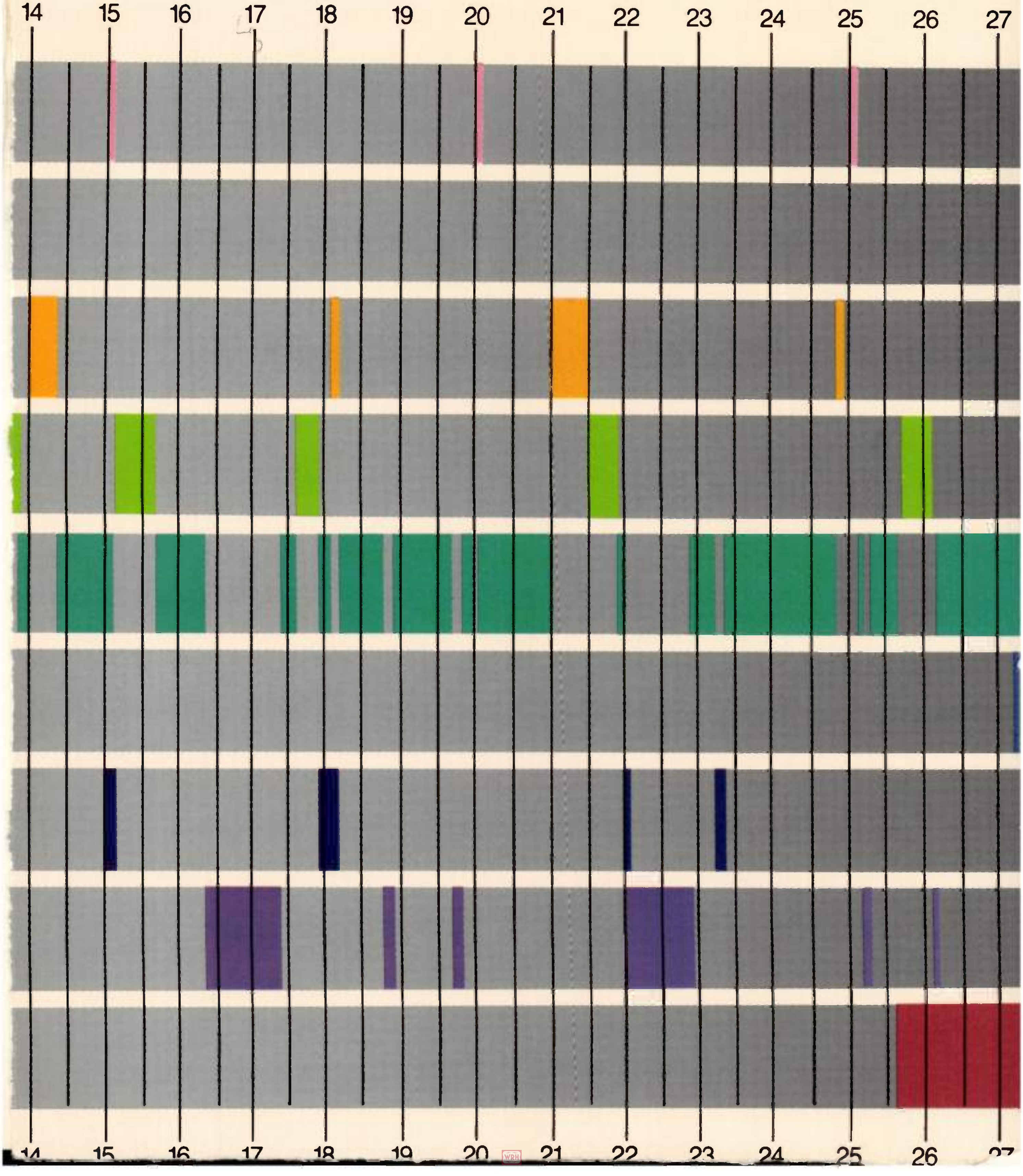

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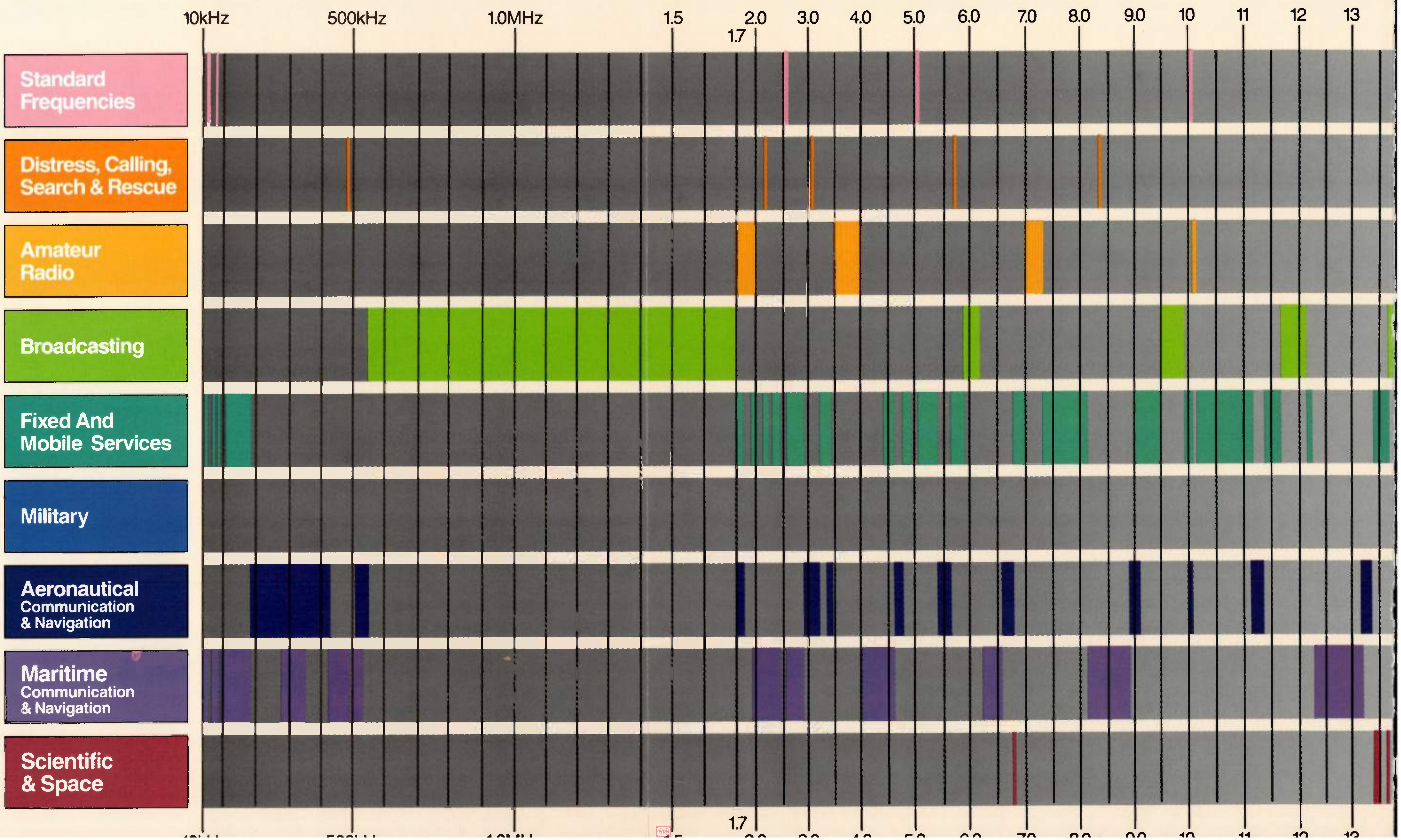

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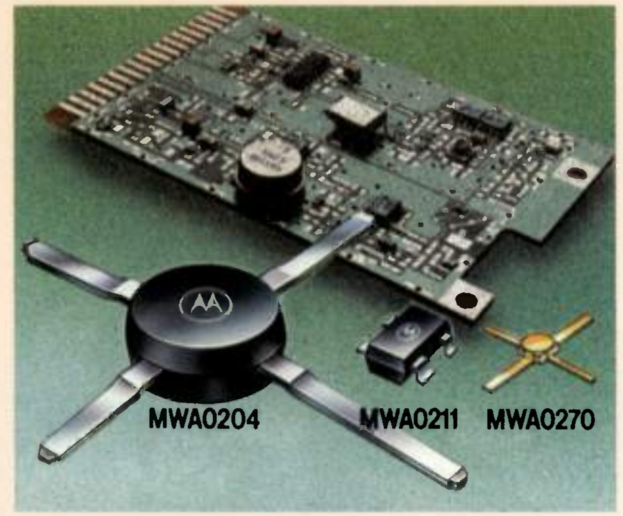

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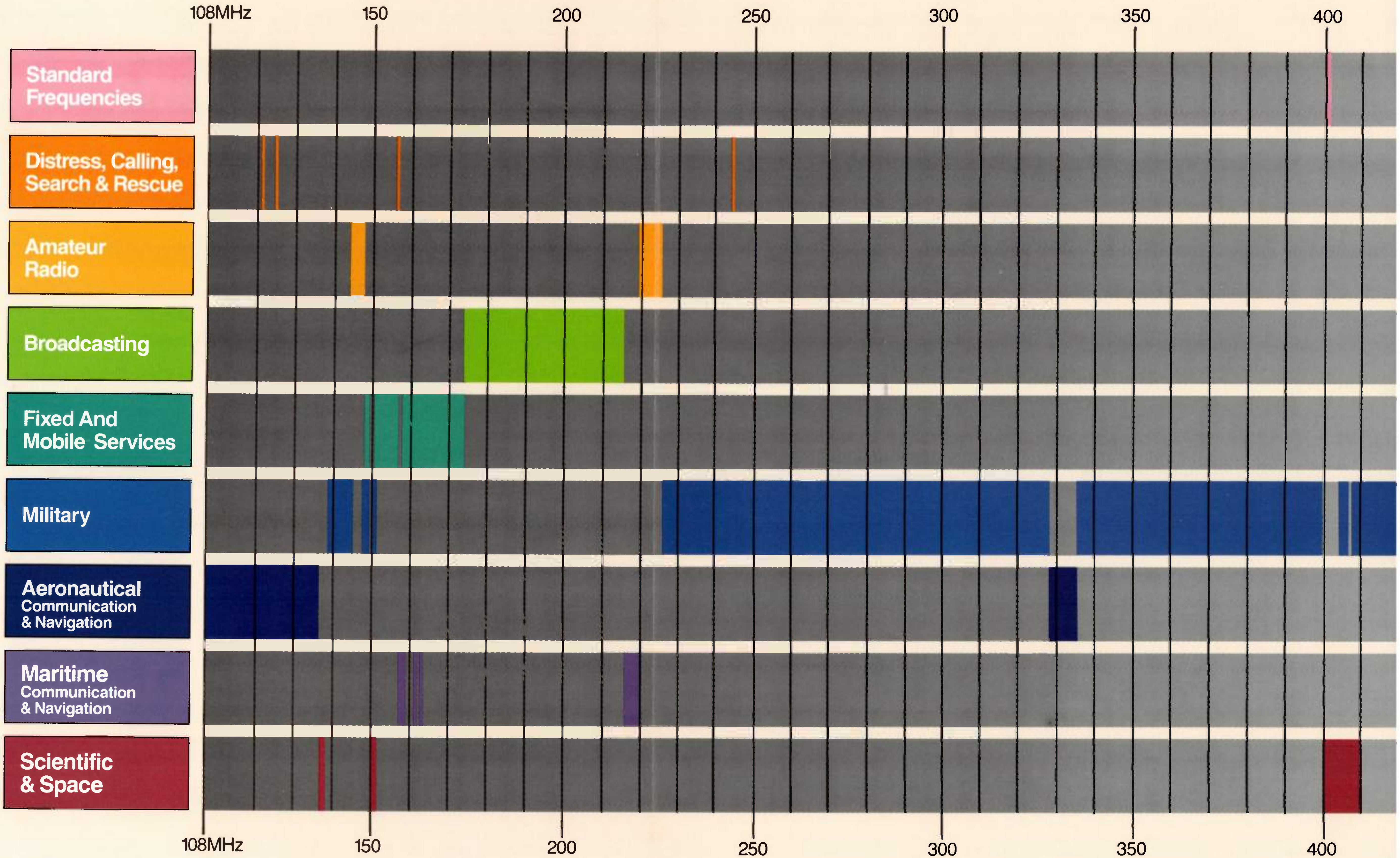
The MRF10120 (120 W) and MRF10030 (30W) help provide the right answers to Mode "S" air-traffic-control transponder questions about location and altitude of planes so collison is avoided. Capable of handling 10 μs pulses at 50% duty cycle for up to 3.5 ms, the units have guaranteed performance including ruggedness. Used at 960 to 1,215 MHz, they have an overall duty cycle of 25% and incorporate input matching for broadbanding, gold metallization, nitride passivation and low thermal resistance for heat dissipation. They are also available for JTIDS use.

High frequency gets broader in 4th edition RF Data Book Selector Guide.

Over 100 new units covering a broad applications spectrum for VHF to 900 MHz have been added to the fourth edition (DL110REV2) of our all-in-one RF Data Book.

Kissed by quietness.

These GHz transistors (MRF941/951 series) whisper at state-of-the-art noise figures and are offered in a variety of packages, plus chips for optimum applications flexibility. Rated at 15, 30, 50 or 100 mA IC, the MRF-MMBR-family provides high-quality, front-end amplification of weak signals in portables or land-mobile radios. Design

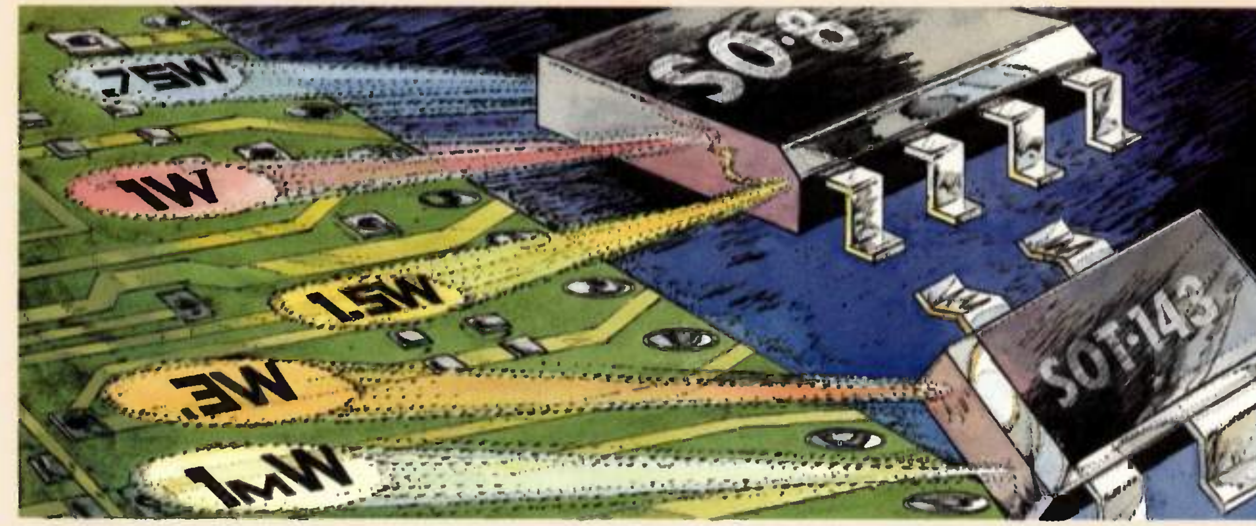


Motorola's RF Semiconductors cov

particular device, the manual includes new, 1.2 dB devices, higher-power (to 600 W) bipolar and TMOS FETs, UHF, 28 V and 900 MHz units, CRT drivers, low-current and VHF parts plus our new, 550 MHz, 6000 series linear modules.

The revised selector guide and cross-reference further simplifies selection with recommended line-ups for a variety of power/frequency applications.

SOT-23/SOT-143 with tape and reel packaging options, or chips where you need all the quiet you can get. Ceramic JAN, JANTX and JANS versions will be offered soon, too.



GHz gain, miniscule drain in SOT-143.

Operating at less than 1 mA I_C with a V_{CC} of only 1 V, the MRF9331 offers 13 dB typical gain at 1 GHz and 2 dB NF at 500 MHz. Use of a single emitter finger of a 1.25 micron die design achieves state-of-the-RF-art performance with high, 6 GHz f_T.

Applications are pagers, portable radios, hearing aids, remote controlled door openers and any equipment operating on batteries you want to last. Wide dynamic range is possible with 100 μA to 2 mA I_C.

And SOT-143s provide dual emitters for best performance in common emitter configurations.

Price: just $1.50, 1K-up.

900 MHz SO-8 grabs 1 W honors.

Providing 1 W and more P_D, equal to or exceeding that of MACRO-X and SOE packaging, the MRF8372 is the SMD version of the popular, 6 GHz f_T MRF837 used in wideband, large-signal, pre-driver, 900 MHz, land-mobile stages.

No sacrifices in technology, either: 2-micron geometry, gold top metal, ion-implanted emitters and silicon nitride passivation.

Typical performance is 10 dB gain at 870 MHz, 12.5 V, in a broadband circuit delivering ¾ W RF power with 60% efficiency. It's also characterized at 7.5 V and at 51 MHz.

Little price, too: $1.65, 1K-up.

Superb 8 GHz performer even more so in SOT-143.

None finer and now in SOT-143. It's the popular MRF571 and it's now MRF5711 in 4-leaded SMD for reduced common-element inductance and higher gain.

Leading-edge die technology, identical to its MACRO-X counterparts, provides 1.7 dB NF and 12 dB gain at 1 GHz.

Available in standard and low profile, this high-performer is ideal for high-gain, low-noise, small signal amplifiers: satellite TV receivers (TVRO), wide dynamic range front end amplifiers and low-noise VCOs. Plus instrumentation applications where high-gain to high-frequencies are required, or in fast switching designs.

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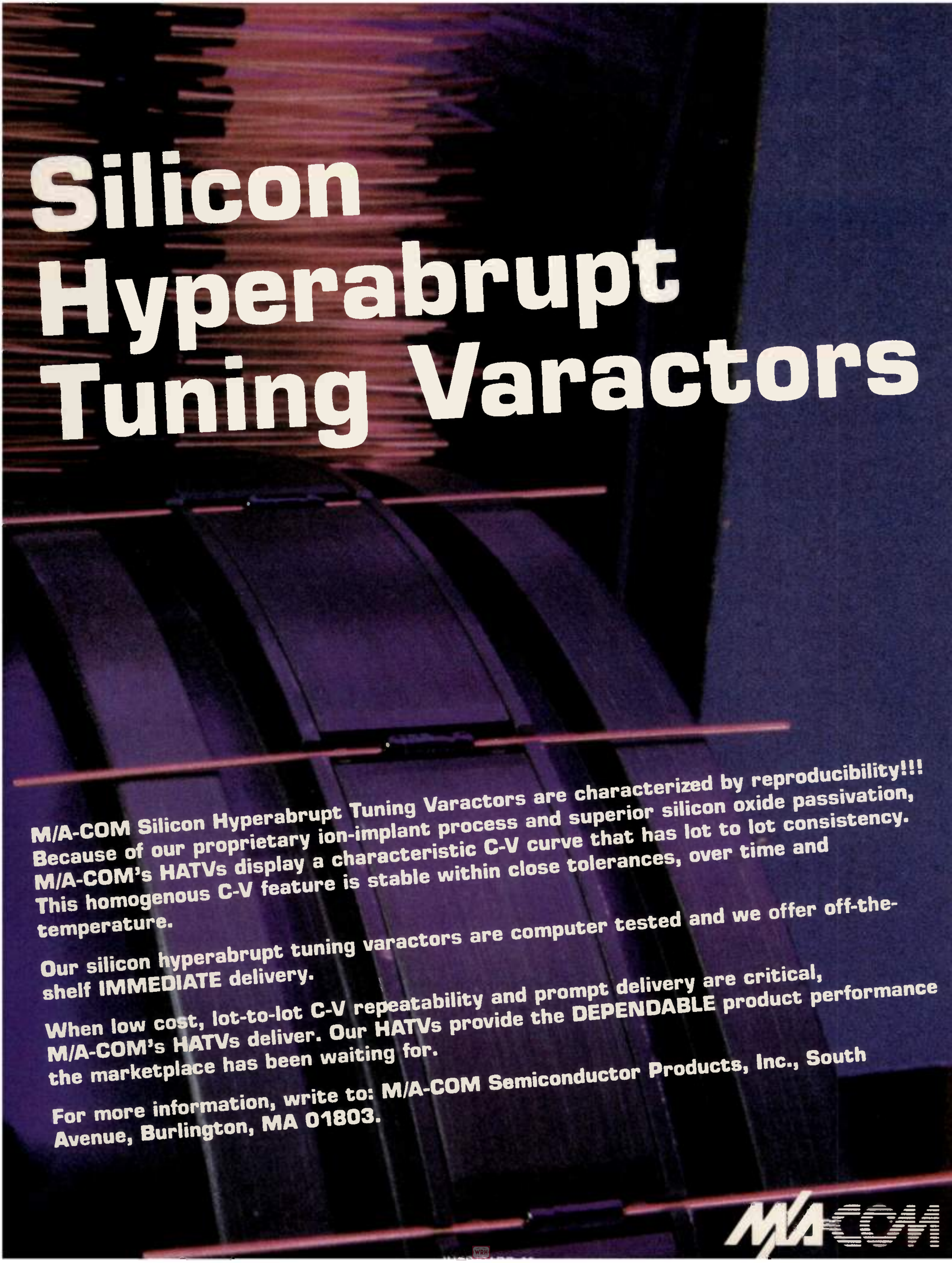
The RF Design/Motorola frequency allocation chart for RF engineers... color-coded to show you bands of frequency allocated to the user in nine categories...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New Integrated Circuits for RF Applications

By Gary A. Breed
Editor

RF engineers can obtain more functionality and better performance than ever in the current generation of RF integrated circuits. A designer now has viable monolithic options for most "generic" circuit functions (preamplifiers, IF amplifiers, mixers, detectors, etc.), in addition to many complex circuits. Since recent developments in GaAs technology have been relatively well publicized, this report will focus on silicon ICs, still the largest part of the integrated circuit industry.

Silicon integrated circuits have been around since the 1960s, and have been the mainstay of both analog and digital devices. Silicon's well-established performance and fabrication technology makes it attractive for the implementation of new integrated functions. Although other technologies offer better performance in some areas, Si offers significant cost advantages and remains a good performer at UHF and lower frequencies.

Amplifiers

The most common integrated circuits for RF and high speed applications are amplifiers. These can be grouped into two main categories: AC-coupled small or medium signal level RF amplifiers; and DC-coupled video amplifiers, operational amplifiers and buffers.

Higher power, lower cost, and greater availability are the latest developments in RF monolithic amplifiers. Avantek's MSA series amplifiers are well known examples, with a wide range of gains and bandwidths. The most recent additions are the MSA-0520, with 8.5 dB gain and a power output of +23 dBm (1 dB compression), and the MSA-1023 with even higher power output of +27 dBm. Both devices are characterized over the 100 MHz-1 GHz range. Linear Technology, Inc., also has a new device with high power, the LB951, with +20 dBm output over 10-200 MHz. The LB951 is fabricated using DMOS technology.

Price and availability have been enhanced with the entry of Motorola and Mini-Circuits Labs into the silicon MMIC amplifier market. Motorola has new MWA02-03 series amplifiers which mimic (pun intended) some of the most popular Avantek devices. Mini-Circuits has been marketing their MAR line of low cost amplifiers for about a year, with prices as low as $.99 for some models (in modest quantities). Other low cost amplifiers include the NE5205 and its lesser bandwidth sibling NE5204 from Signetics.

In the DC-coupled arena, where applications range from PLL loop amplifiers to data acquisition signal conditioning to video baseband systems, there is plenty of activity. Comlinear Corp. has just introduced its first monolithic amplifiers, models CLC400 and CLC401. This company's expertise in high speed electronics has previously been limited to hybrid products. The lower cost of monolithic fabrication ($15.50 in 100s, plastic package) makes these high performance amplifiers attractive for new applications.

High speed operational amplifiers are available from many IC manufacturers. Two recent entries are the PA19 power op amp and the WA01 wideband amplifier from Apex Microtechnology Corp. The PA19 can deliver up to 60 watts RMS with a gain-bandwidth product of 100 MHz. The WA01 is a 4-watt amplifier with a 4000 V/us slew rate and 150 MHz power bandwidth. Plessey has the SL2541 op amp, a device with a unity gain bandwidth of 800 MHz and programmable open loop gain, supply voltage range and output current and DC offset. VTC Inc., a Minnesota IC maker, has their VA700 line of high speed/precision op amps for data conversion systems or signal processing applications. Harris Semiconductor, Signetics, Burr-Brown, Texas Instruments Analog Devices, GE/RCA and National Semiconductor have for some time offered high speed op amps in a variety of performance ranges.

Video amplifiers are another class of high speed products. The two major applications are baseband amplification (true video or other wideband signals) and magnetic storage media read amplifiers. Recent entries include the MAX452 family from Maxim, which include on-chip multiplexers, and the LH4002 buffer/current driver from National.

Switches

Monolithic analog switches are achieving the switching speeds and low insertion loss necessary to replace diode switching in many RF applications. High frequency choppers, modulators, mixers and multiplexers are typical applications, as well as signal path switching. Leading the way in high performance is DMOS technology, often combined with CMOS driver circuitry. Topaz Semiconductor and Siliconix Inc. are leaders in this area, with a wide range of DMOS switch and transistor products.

PIN diodes are still the mainstays of RF switching, with unique drive requirements. Although many "stock" logic devices can be utilized as PIN drivers, some ICs have been designed with this application in mind. Intersil has a relatively new part, the ICL7667 dual power driver, which has been characterized for use as a PIN driver.

Multi-Function ICs

One benefit of improving fabrication techniques is the ability to put several functions on one chip. Last January, we featured an FM IF integrated circuit on our cover, the Motorola MC3362, with two mixer/local oscillator blocks, a limiting IF amplifier, signal strength indicator (RSSI), quadrature detector, audio preamp and a comparator for data recovery. Motorola has now introduced the next version of the IC, the MC3363, with RF amplification and squelch.

Signetics has made a similar move toward greater integration, with the NE605, a single IC combining the functions of their NE602 mixer/oscillator and NE604 FM IF strip, with limiting IF amplifier, RSSI, quad detector and audio preamp. This chip has all circuitry (except audio power amplification) for a single-conversion FM receiver.

Another company involved in receiver ICs is Siltronics Ltd., with products for FM receivers and low cost paging applications. Like others in this market area, they emphasize low voltage operation and low power consumption for simple battery-operated receivers.

Signetics, Motorola, Plessey and National Semiconductor have many other receiver ICs for consumer AM, FM and TV receivers, remote controls, wireless intercoms, and other RF functions. These lines have developed over many years, and a review of their catalogs and data sheets offers a long list of inexpensive RF ICs.

ASICs (Application Specific Integrated Circuits) have been the fastest growing part of the digital IC world, and are just getting started in high frequency analog technology. The main hurdle in analog ASICs has been the development of fast enough PNP transistors to utilize existing fabrication techniques at high frequencies. At lower frequencies, where "standard" op amps and comparators can be used, companies such as Raytheon have has ASICs available for some time.

At frequencies high enough to be considered RF, VTC Inc. has solved the PNP transistor problem to achieve up to 500 MHz bandwidth in bipolar linear and mixed linear/digital products. For example, their VL3000 bipolar cell library can include amplifiers, comparators, DACs, line drivers and VCOs plus many TLL or 10K ECL digital functions in such applications as telecommunications, signal processing, local area network systems or instrumentation.

Digital/RF Circuits

Digital integrated circuits for RF applications include both familiar and new functions and capabilities. Most familiar are the prescalers and dividers for PLLs or frequency counters. Manufacturers have been pursuing GHz speed performance in silicon with dividers operating up to about 3 GHz.

Complete PLL circuits on a single chip have reached new levels of performance in low power, high speed, and comprehensive functionality. Motorola has low power CMOS ICs (MC145159 and others) which operate in the low MHz range and use external prescalers for higher frequencies, while Signetics and Plessey are approaching 1+ GHz performance on a single chip using bipolar processes (external VCO).

Direct digital frequency synthesis (DDS) is a rapidly growing part of digital/RF development. Sciteq has an ECL device developed specifically for phase accumulator applications in DDS. Stanford Telecommunications has the recently upgraded ST-1172A, a nearly complete (less clock and DAC) direct digital synthesizer IC for generating output signals

Avantek, Inc., Santa Clara, CA. INFO/CARD #136.
Linear Technology, Inc., Burlington, Ontario. INFO CARD #135.
Motorola, Inc., Bipolar Analog IC Division, Tempe, AZ. INFO/CARD #134.
Mini-Circuits Laboratories, Brooklyn, NY. INFO/CARD #133.
Signetics Corp., Sunnyvale, CA. INFO/CARD #132.
Comlinear Corp., Ft. Collins, CO. INFO/CARD #131.
Apex Microtechnology Corp., Tucson, AZ. INFO/CARD #130.
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Harris Semiconductor, Melbourne, FL. INFO/CARD #127.
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Burr-Brown, Tucson, AZ. INFO/CARD #125.
Texas Instruments Inc., Dallas, TX. INFO/CARD #124.
GE Solid State, Somerville, NJ. INFO/CARD #123.

up to about 10 MHz. Newer still is their STEL-2172 ECL device with similar architecture, but a 300 MHz clock rate for up to 130 MHz frequency synthesis. Also in the DDS realm is Digital RF Solutions' DRFS™-2250 and -3250 synthesizers, and accessory chip set with complete digital frequency, phase and amplitude modulation capability.

Digital signal processing is a term that encompasses a wide range of RF applications, both in communications and instrumentation. The primary devices in these applications are the analog-to-digital and digital-to-analog converters (ADCs and DACs). New ADC and DAC products are announced almost daily. Some of the latest include the AD9002 150 MHz sample rate 8-bit ADC from Analog Devices. The TDC1318 triple 8-bit DAC from TRW LSI Products with 200 MHz operation, and the MAX162 from Maxim, a 12-bit CMOS successive approximation ADC with 4 MHz clock rate.

Operation up to several hundred MHz is available in monolithic DACs (8-bit) from TRW, Analog Devices, Burr-Brown, Honeywell, Intech, Sony, Toshiba and others. These same companies offer high speed ADCs with sample rates up to 150 MHz for 8-bit devices, 20+ MHz for 12-bit precision.

Among digital signal processing ICs recently introduced is the PDSP16330, dubbed by manufacturer Plessey as a Pythagoras Processor. This LSI circuit is a cartesian to polar converter, accepting 16×16 bit magnitude data and outputting 16 bit magnitude and 12 bit phase information.

The military-sponsored VHSIC (Very High Speed Integrated Circuits) program of 1.2 and 0.7 micron CMOS technology has begun to generate "real" products, although currently limited to military projects. As the technology matures, commercial applications, primarily in digital signal processing, will certainly appear.

This brief overview of progress in silicon ICs for RF and other high speed/high frequency applications can only serve to whet our appetites for more information. Checking the Info/Card numbers below will bring additional information on IC products. rf

National Semiconductor, Santa Clara, CA. INFO/CARD #122.
Maxim Integrated Products, Sunnyvale, CA. INFO/CARD #121.
Topaz Semiconductor, San Jose, CA. INFO/CARD #120.
Siliconix, Inc., Santa Clara, CA. INFO/CARD #119.
Intersil Inc., Cupertino, CA. INFO/CARD #118.
Siltronics Ltd., Kanata, Ontario. INFO/CARD #117.
Raytheon Co.,Semiconductor Div., Mountain View, CA. INFO/CARD #116.
Sciteq Electronics Inc., San Diego, CA. INFO/CARD #115.
Stanford Telecommunications, Santa Clara, CA. INFO/CARD #114.
Digital RF Solutions, Santa Clara, CA. INFO/CARD #113.
TRW LSI Products, La Jolla, CA. INFO/CARD #112.
Honeywell Inc., Signal Processing Div., Colorado Springs, CO. INFO/CARD #111.
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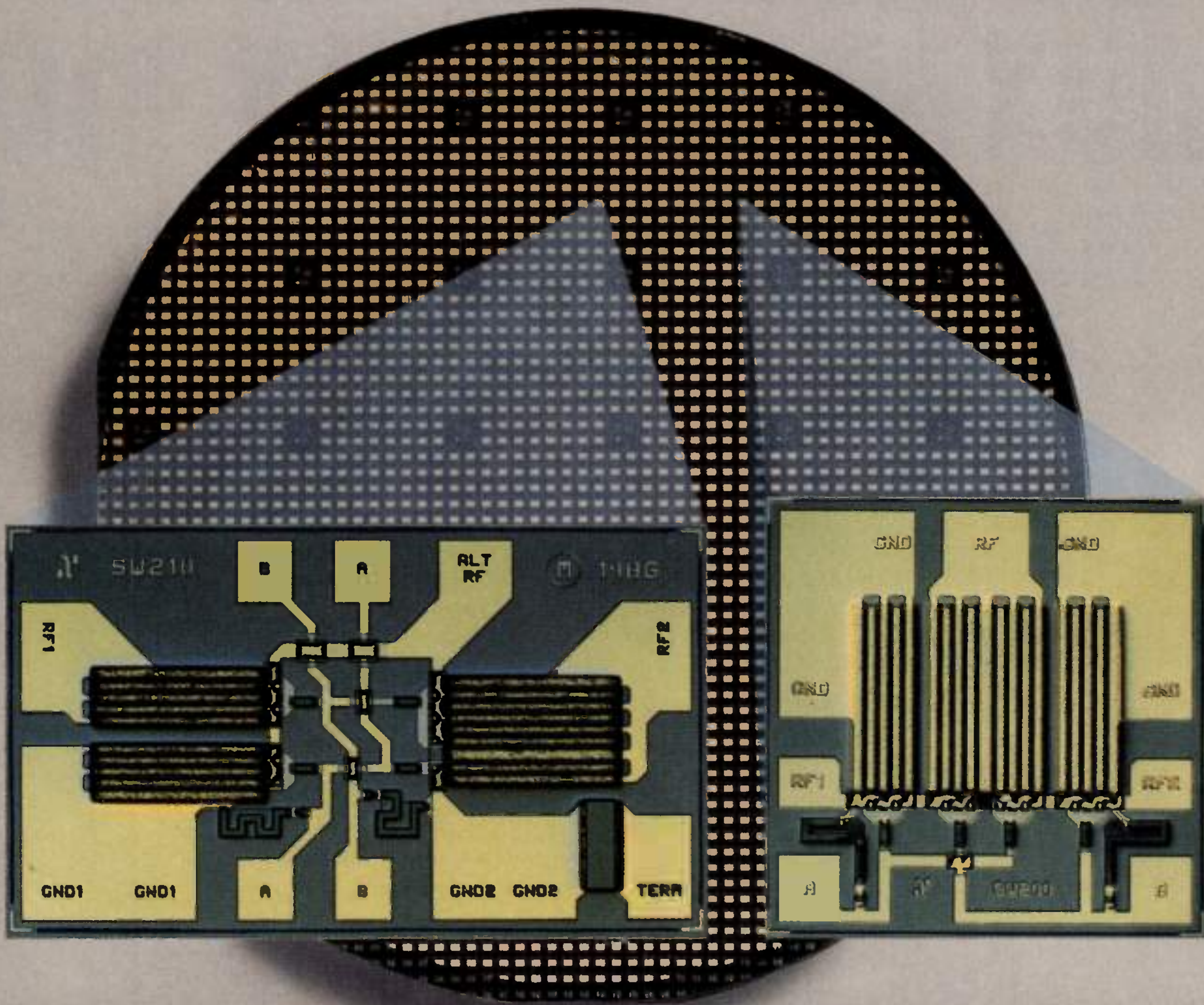

GaAs MMIC SPST Configurable Switch
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Frequency Range	DC to 6 GHz
Insertion Loss	0.6 dB max
Isolation	55 dB min
Switching Speed	3 ns typ
VSWR	1.15:1 max
Control Voltages	
V_{IN} Low	0 to −0.2V @ 20 μA max
V_{IN} High	−5V @ 70 μA typ
Input Power for 1 dB Compression	27 dBm typ
Die Size	0.030″ × 0.050″ × 0.010″

GaAs MMIC SPDT Switch
Model SW-200

Frequency Range	DC to 6 GHz
Insertion Loss	0.6 dB max
Isolation	45 dB min
Switching Speed	2 ns typ
VSWR	1.2:1 max
Control Voltages	
V_{IN} Low	0 to − 0.2V @ 20 μA max
V_{IN} High	−5V @ 40 μA typ
Input Pwr 1 dB Comp	+ 25 dBm typ
Die Size	0.030″ × 0.030″ × 0.010″

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Designing Effective Two-Tone Intermodulation Distortion Test Systems

By Manfred Bartz
Hewlett-Packard Company

A firm grasp of the fundamentals of intermodulation distortion (IMD) is key to designing receivers for distortion dynamic range. Of importance is their application to the design of two-tone intermod test systems for verification of receiver performance. Assuring the integrity of intermod measurements requires careful evaluation of the test system margins. A quantitative method for predicting system IMD performance is derived and techniques for extending test system dynamic range are demonstrated.

The transfer characteristic of nonlinear memoryless networks can be represented by a power series expanded about the operating point of the network as follows:

$$V_o = a_o + a_1V_{in} + a_2V_{in}^2 + a_3V_{in}^3 + \ldots \ldots + a_nV_{in}^n \quad (1)$$

This expression can be derived analytically or approximated by using empirical methods for obtaining the polynomial of best fit. The utility of the power series expansion resides in establishing the relationship between the network transfer characteristic and its predicted two-tone intermodulation performance. Setting the input stimulus V_{in} equal to two sine waves of different frequencies models the two-tone test condition as shown in equation 2.

$$V_{in} = V_1\cos(\omega_1 t) + V_2\cos(\omega_2 t) \quad (2)$$

Distortion product terms occur at frequencies which are the sums and differences of multiples of the original two-tone inputs.

$$|\pm n \cdot f_1 \pm m \cdot f_2| \text{ n,m integer} \quad (3)$$

These are depicted in Figure 1. The order of each distortion product is defined to be (n + m).

Although higher order terms in the

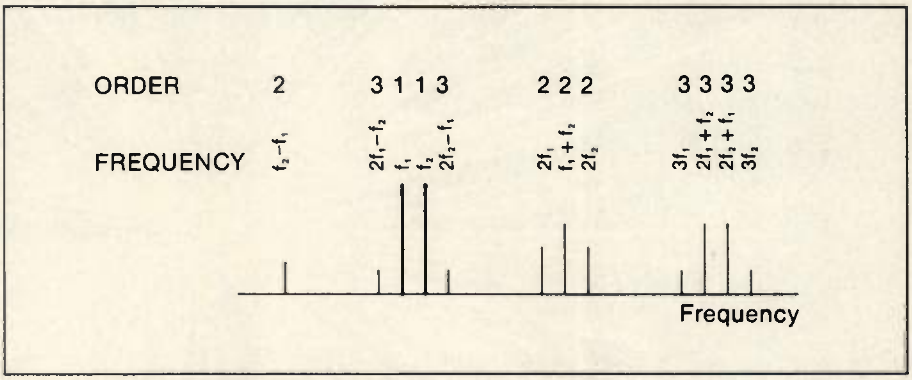

Figure 1. Second and Third Order Two-Tone Distortion Products.

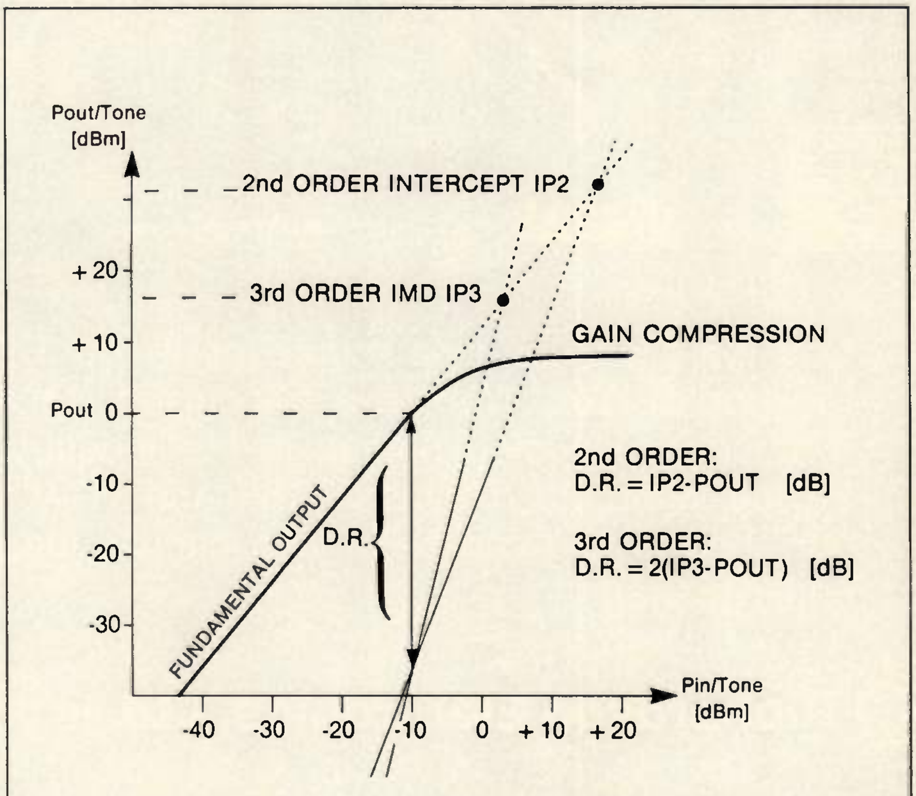

Figure 2. Distortion Dynamic Range Plot Depicting Intercept Points.

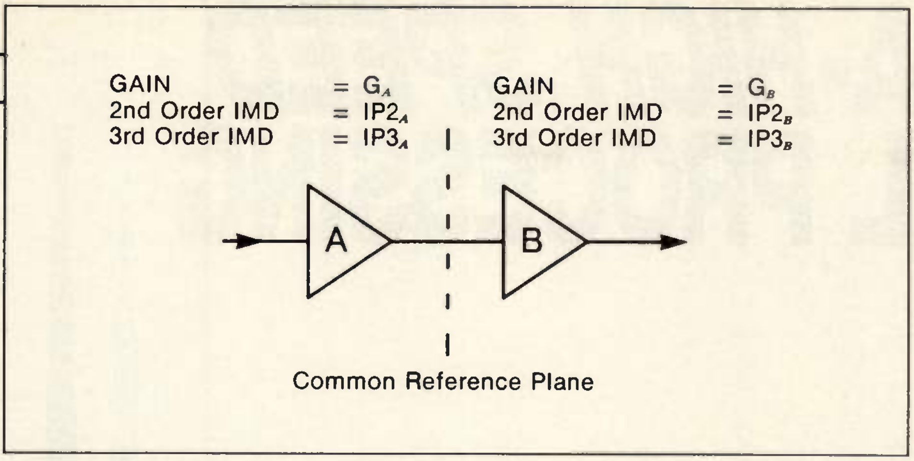

Figure 3. Determining System IMD For Amplifier Pair in Cascade.

power series can contribute to the magnitude of lower order distortion products, for small signal inputs the contribution becomes negligible. Hence it can be shown that the second order intermod is generated by the squared term in the power series and similarly the third order term. For small signals the product of their magnitudes tend toward zero. This is known as the rule of gradual nonlinearity.

The coefficients of the distortion terms are products of the original amplitudes of the two test tones and are also of the same order as the corresponding frequency term. This means a 2-1 distortion product (third order) could have any third order combination of the original amplitudes, e.g. a_1^3, a_2^3, $a_1a_2^3$ and $a_2^2a_1$. From this, one obtains the useful property for identifying the order of a particular distortion term during a measurement. If the amplitudes of both the two tones are dropped by 1 dB, then for a second order term the distortion will drop by 2 dB and the third order term will drop by 3 dB. Therefore, relative to the magnitudes of the original test tones, second order distortion will drop 1 dB per dB and third order distortion will change 2 dB per dB.

Assume a test tone pair is equal in signal level. Then the relationship between second and third order IMD levels to input power level is given by the graph of output power per tone versus input power per tone in Figure 2. Since the distortion dynamic range is a function of absolute input power level, the concept of intercept point is an appropriate convention for expressing distortion figure of merit in a receiver. The second and third order intercept points are the theoretical power levels at which the distortion products become as large as the fundamental power levels. These points are not directly measured but extrapolated, since most receiver components such as mixers and amplifiers exhibit practical operating limits with the onset of gain compression.

For large input signal levels, the higher order terms in the power series expansion of the transfer characteristic become more significant and can interfere with the results. Hence a receiver should always be tested within its linear range. Incidentally, when generating IMD plots for components using the two tone approach, it is reasonable to plot the input power level as 6 dB above the signal level of each of the two tones. This follows because the distortion mechanisms result from peak voltage excursions. The peak envelope voltage of the two equi-level tones is 6 dB above each test tone, or 3 dB above the combined average power of the two tones.

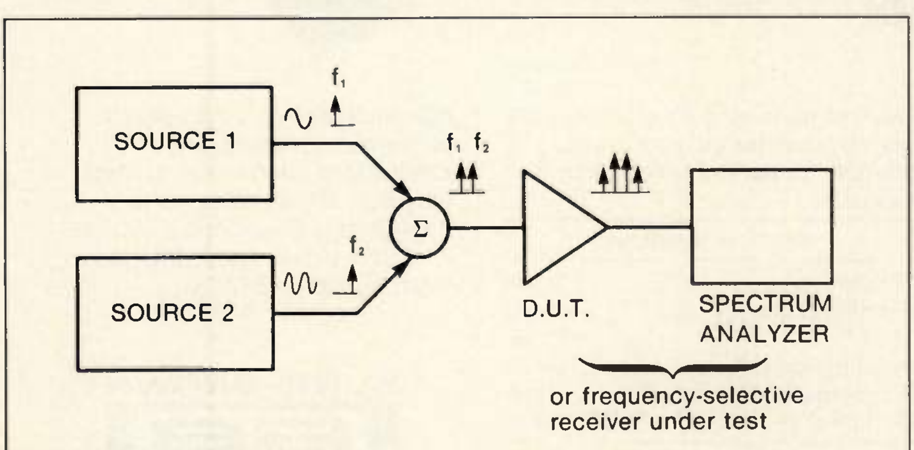

Figure 4. Basic Two-Tone Test System.

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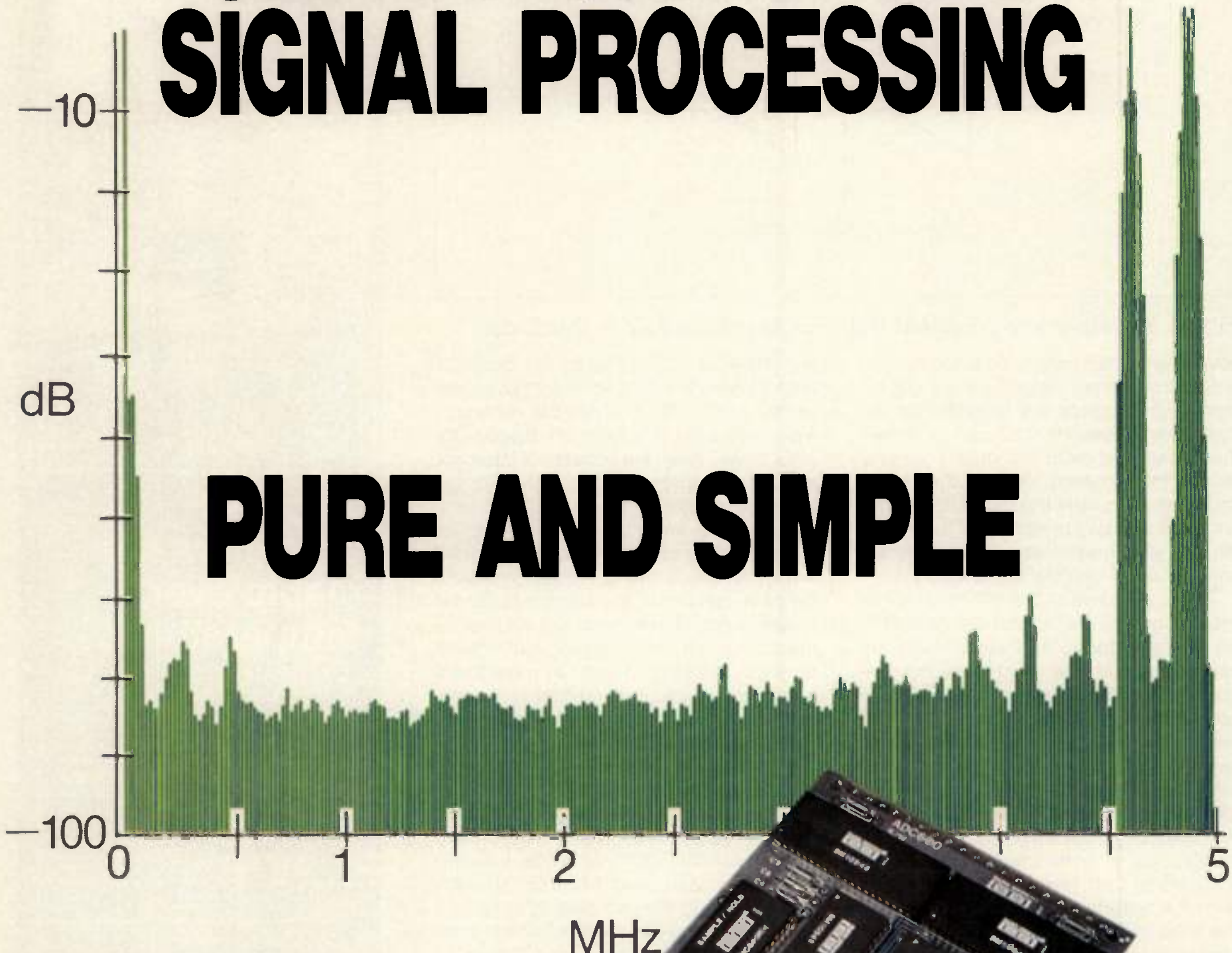

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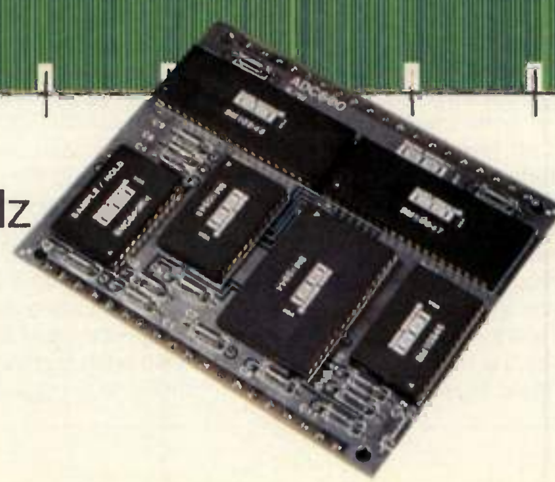

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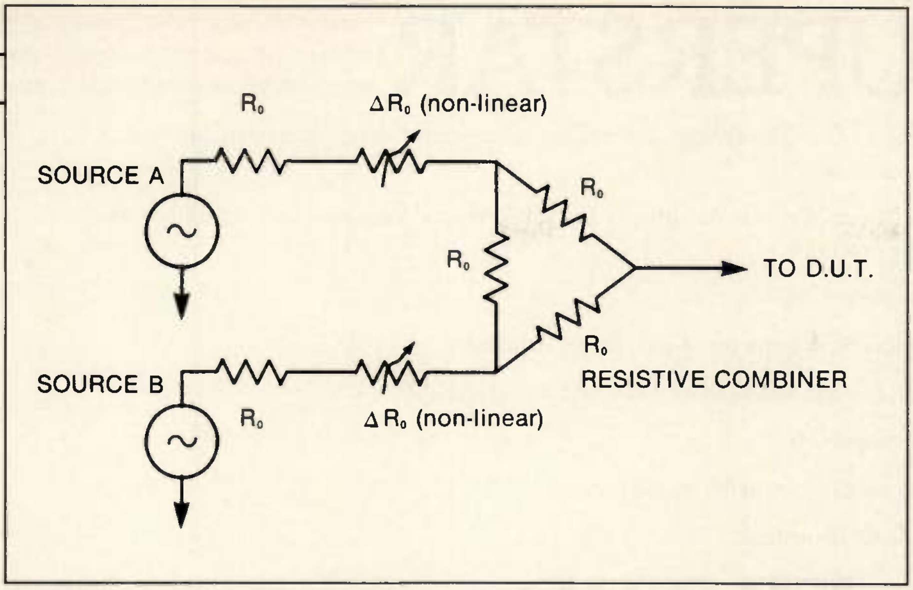

Figure 5. Circuit Model of Combined Two-Source Intermodulation Effect.

Calculating System IMD

The concept of intercept point lends itself to evaluating the overall distortion performance of a system or cascade of blocks whose intercept points are known. The RF designer, who has analyzed or measured the intermod performance of the individual components or sub-blocks of a receiver, can determine the optimum block diagram configuration for achieving a specified dynamic range. This should be carried out in conjunction with a noise figure analysis to assure receiver sensitivity. The technique can also be extended to quantify the integrity of the intermod distortion test system used to evaluate a receiver's performance.

Consider two IF amplifiers A and B in cascade shown in Figure 3. With measured third order intercept points of $IP3_A$ and $IP3_B$ and gains G_A and G_B at the IF frequency, how do we determine the combined intercept point of the two amplifiers? There are many ways in which distortion

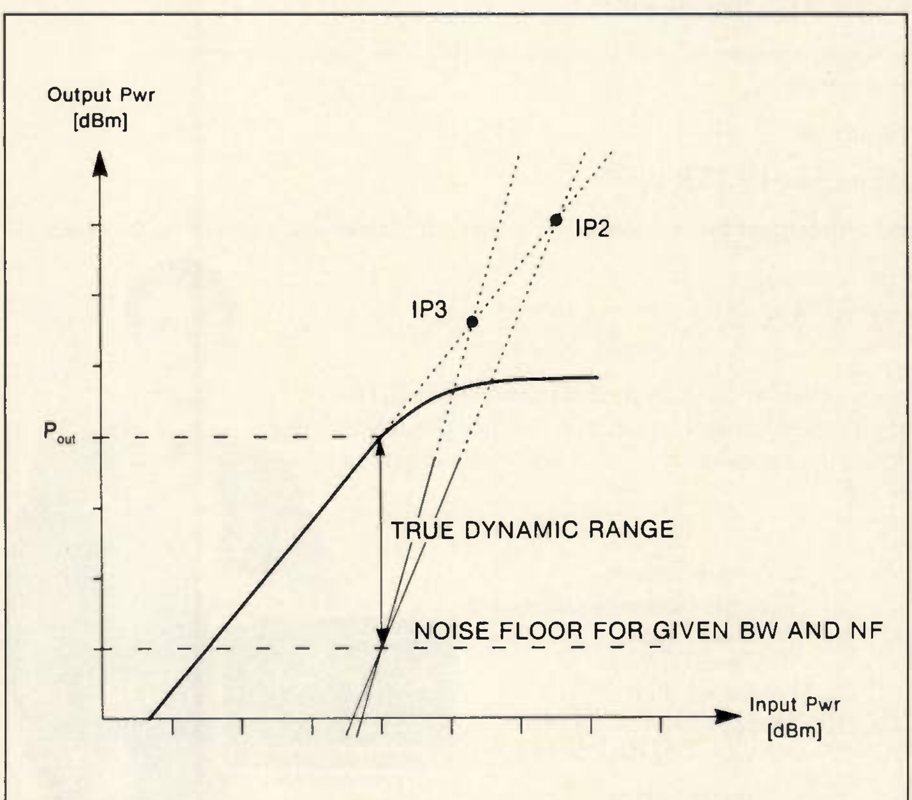

Figure 6. Complete Dynamic Range Plot with Noise Floor.

SUPERSTAR

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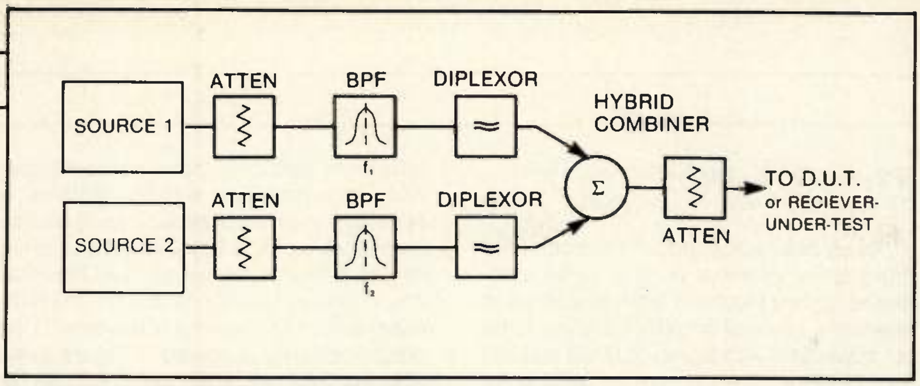

Figure 7. Enhanced IMD Test System.

products can add depending on whether their power levels are correlated and the nature of their phase relationship. These effects can result in a worst case scenario of direct voltage addition of products in-phase or even complete cancellation. By assuming the products result from direct in-phase algebraic addition of voltage levels, the designer assures a worst case margin which is reasonably accurate in practice.

Choosing a common reference plane between the two amplifiers, the intercept point for amplifier A must be referred to the output and the intercept point for amplifier B to its input. This normalization is performed by adding or subtracting the amplifier gain in dB from its respective intercept point in dBm. In addition, input and output power levels must be referred to the common reference plane. From the intermod plot it is evident that the intermod products for third order IM can be computed in dB below the two-tone signal levels as follows:

$$IM3 = P_o - 2\,(IP3 - P_o) \qquad [dBm] \quad (4a)$$

Taking the antilog of (4a) yields the following equivalent expression:

$$IM3 = P_o^3/IP3^2 \qquad [mW] \quad (4b)$$

Substituting $V = (P_o.R)^{1/2}$ converts the expression from units of power to units of Volts.

$$VIM3 = \left[\frac{(P_o)^{3/2}}{IP3}\right] R^{1/2} \qquad [V] \quad (5)$$

Since we postulated that the intermod product voltages from each amplifier add algebraically, we find that:

$$VIM3_T = VIM3_A + VIM3_B \qquad [V] \qquad (6)$$

Substituting previous equations and applying (4b) results in the solution for the total intermod as

$$VIP3_T = \left[\frac{1}{VIP3_A} + \frac{1}{VIP3_B}\right]^{-1} \quad [V] \qquad (7)$$

In dB, this yields the following expression:

$$IP3_T = -10 \text{ Log } [10^{-IP3_A/10} + 10^{-IP3_B/10}] \quad [dBm] \qquad (8)$$

In many measurement situations, not limited to narrowband receivers or I.F. subsystems, second order intermod becomes equally important. An example would be a wideband receiver front end. By similar analysis it can be shown that linear and logarithmic results for second order combined intercept are given by the following equations.

$$VIP2_T = \left[\frac{1}{(VIP2_A)^{1/2}} + \frac{1}{(VIP2_B)^{1/2}}\right]^{-1} [V] \quad (9)$$

$$IP2_T = -20 \text{ Log } [10^{-IP2_A/10} + 10^{-IP2_B/20}] \quad [dBm] \qquad (10)$$

As an example, suppose the combined third order intercept point is to be computed for two Motorola MSA amplifiers in cascade. The first amplifier is given to be an MSA-0204 with a gain of 17 dB and IP3 of +15 dBm. The second, an MSA-0404 has a gain of 8 dB and IP3 of +26 dBm. Both intercepts points are referred to the output. Choosing the reference plane to be between the two amplifiers, the output intercept point of the second amplifier, referred to its input by 8 dB of gain, yields a value of +18 dBm. Substituting these values into equation 5 yields a combined intercept of +13 dBm at the plane of reference or +12 dBm referred to the output of the cascade.

IMD Test Systems

The calculation for system IMD applies to evaluating the performance of intermodulation distortion test systems as well. The most elementary block diagram of a two-tone test system depicted in Figure 4 consists of two sources connected by a resistive combiner. Unless the device under test (DUT) itself is a receiver with frequency selective measurement capability, the complete system requires a spectrum analyzer. Although each source generates a single tone, intermodulation occurs between the two sources because each source mutually modulates the nonlinear output impedance of the other. This phenomenon is modeled in Figure 5. An effective intermod intercept point can be assigned to the sources at a given frequency or valid range of frequencies. If the two sources are of different model types, then the IMD contribution from each will be different as well. This can be measured for each source and computed with the previous expressions or simply measured for the two-source combination.

Given that a certain IMD performance will be required of the receiver under test, how does one determine the required IMD margin for the two sources? Using the case of third order intermod as an example and equation (8), one can easily verify the following: Given that the test system has a system IP3 that is 10 dB higher than the DUT, the system can contribute a measurement error as large as ±2.5 dB. For 20 dB of test system margin, the uncertainty of the measurement is reduced to ±0.8 dB worst case.

If the DUT is not a frequency-selective receiver but rather a receiver sub-block or component, a spectrum analyzer becomes necessary to measure the intermodulation products. Again, the previous technique can be used to evaluate the test system intermod design margins to account for the analyzer intercept. However, additional factors limiting the dynamic range of the measurement must be taken into consideration.

Most spectrum analyzers have a front-end attenuator to assure the input signals fall within the linear range of the input mixer. If the front-end attenuator proves insufficient, an attenuator can be added to the test system block diagram. As previously indicated in Figure 2, by increasing attenuation at the front-end of the analyzer, one can achieve an input signal level which will guarantee the needed distortion dynamic range in the analyzer. However, now the noise floor sensitivity of the analyzer comes into play. It is entirely possible to discover for the optimum levels chosen that the intermod products to be measured fall below the noise floor of the analyzer. Since the noise floor is bandwidth dependent, the true dynamic range of the analyzer can be computed if the noise figure, intercept point and bandwidth of the analyzer are known. This is given by expression (11) and depicted on the dynamic range plot in Figure 6.

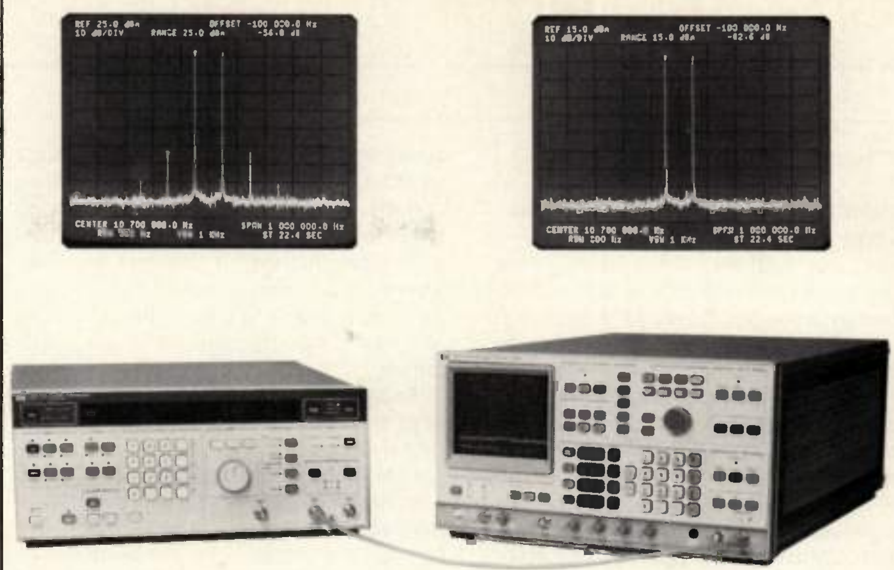

Figure 8. Demonstration of attenuator isolation technique for reducing source intermodulation effects using HP3326A dual channel synthesizer and HP3585A spectrum analyzer.

Noise D.R. =

$$P_o - (174 - 10 \text{ Log BW} - \text{N.F.})[\text{dB}] \quad (11)$$

In addition to adequate distortion intercept test margins and noise floor of the test instrumentation, the gain or insertion loss of the DUT must be taken into account in choosing optimum test signal levels for the test system. These are the parameters the designer must consider in determining the measurement window for optimum results.

Enhancing Test System Dynamic Range

For applications demanding wide dynamic range, the design of high integrity test systems begins with the selection of low distortion sources and a wide dynamic range analyzer. In cases where the measurement application exceeds the capability of the best instruments or what is available, there are techniques for extending dynamic range limitations of the test system. The block diagram of an enhanced IMD test system is shown in Figure 7. Most of these approaches center on increasing the isolation between the sources. The proper choice of combiner for the system can substantially increase the isolation while simultaneously reducing its insertion loss. In Figure 4 the presumed combiner was resistive in nature providing impedance matching at all three ports. Such a combiner has the advantage of working at all frequencies down to DC. However, it has an insertion loss of 6 dB and the isolation it provides between the sources is minimal.

For narrowband applications the use of a hybrid transformer can improve the isolation between the sources by as much as 20 dB or more with an insertion loss of 3 dB. The 20 dB of isolation translates direct improvement to the combined intercept performance of the two sources. There are a variety of hybrid transformers such as the zero degree hybrid or magic tee which can be tailored to the application.

Another means for assuring the integrity of the test system is to insert lowpass or bandpass filtering in each arm of the sources. The purpose of these filters is to eliminate the possibility of interference from unwanted spurious or harmonic frequencies generated by the sources. At VHF or UHF frequencies the additional use of diplexors between the filter elements is recommended to reduce the effects of out-of-band impedance mismatch on IMD performance. When using filters in the audio range take caution in utilizing iron core inductors. These can generate distortion with the onset of core saturation because of the magnetic nonlinearity of their iron cores.

Although these approaches can improve the test system, the tradeoff along with additional complexity includes more constraint on the frequency range of the test system. For narrowband IF testing this may not be a hindrance at all; however, in broadband receiver testing, in which many test points may be necessary across the band both for second and third order intermods, this may require several test configurations.

Additional broadband isolation may be achieved from an inherent feature of many sources that have an output step attenuator in their block diagram. For lower signal levels which cover the majority of receiver related or amplifier testing, additional isolation is obtained from the output step attenuator itself. For this, the source block diagram must have the output step attenuator located between the output amplifier and the source output.

One source particularly suitable for intermod testing of receiver IFs and linear amplifiers to 13 MHz is the HP3326A dual channel synthesizer. It features independent frequency and amplitude control for each channel and has a switchable internal combiner. This eliminates the need for two sources. The internal block diagram of each channel exemplifies the attenuator isolation technique. Although the intermod specifications for IMD are quite respectable for full power output, these improve directly for increasingly lower output signal levels with each step attenuator range in each channel. With an output level of 17.96 dBm at 455 kHz the IMD is 80 dB. At the first step attenuator range the output is 7.96 dBm and thereby provides 20 dB additional isolation between the channels. This results in a 20 dB improvement in third order intermod.

Because of its wide dynamic range for both distortion and noise, the HP3585 is an appropriate companion to the HP3326A for IMD test systems in this frequency range. The plots of a two-tone test measured with the HP3585A spectrum analyzer for a 10.7 MHz IF are shown in Figure 8. With signal levels generated by the HP3326A at both full power and at the first attenuator setting Figure 8 demonstrates the marked improvement that attenuator isolation can provide in lower signal level applications.

Through application of the previous techniques, it is often possible to achieve intermod performance from the sources which yield intermod products well below the IMD performance of the DUT. This is desirable for signals applied to the input of the DUT, but often the IMD products generated by the DUT still exceed the dynamic range of the analyzer. In this case the use of over driving techniques are warranted. By raising the signal levels at the DUT sufficiently, such that the distortion is dominantly generated by the DUT, the intermod products can be prominently measured by the analyzer. Using the dynamic range plot, the actual intermod dynamic range is extrapolated below the noise floor from the data points. The guiding constraint in using this approach is to ensure that the input signal levels are well within the linear range of the DUT. Keeping in mind that for large level input signals, higher order terms can contribute to the second and third order results.

In summary, the proper application of intermod fundamentals is the key to good receiver design. Moreover, the measurement integrity of the test system used to verify receiver performance must be even better and should allow sufficient test margins. The extension of the concept of intercept point to include systems yields a quantifiable approach to determining a figure of merit for evaluating intermod distortion test systems. This methodology allows one to calculate those margins with confidence because the derivation is based on worst case assumptions. rf

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About the Author

Manfred Bartz is an RF design engineer at Hewlett-Packard Company, Lake Stevens Instrument Division, 8600 Soper Hill Road, Everett, WA 98205-1298. He holds a BSEE from the University of Colorado in Boulder and an MSEE from the University of California, Berkeley. He can be reached by telephone at (206) 335-2241. His Ham call is N7GBL.

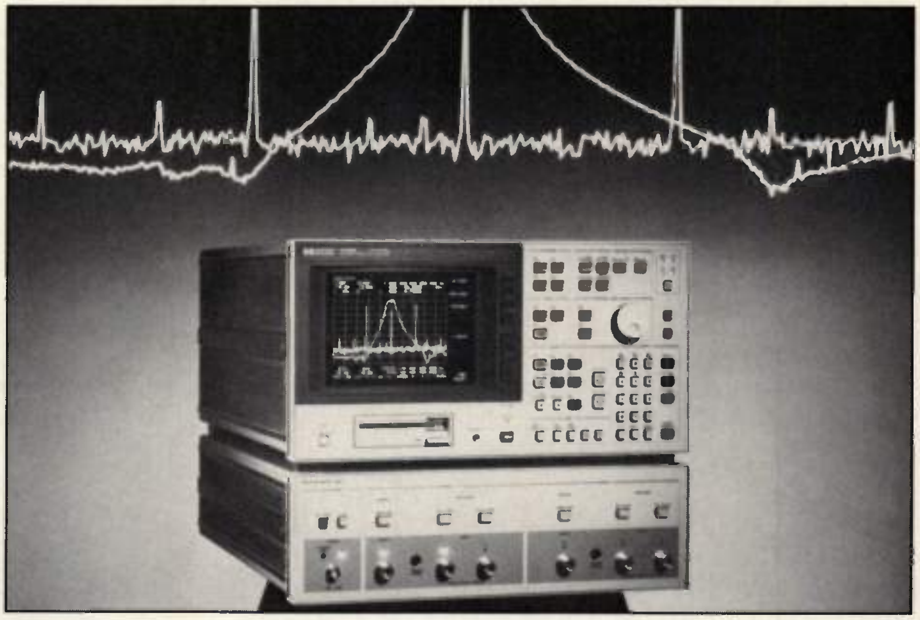
New Products Featured at RF Expo East

Hewlett-Packard Chooses Boston to Unveil Spectrum/Network Analyzer in the U.S.

The HP 4195A network/spectrum analyzer is an instrument that provides vector network and spectrum measurements in one unit with a frequency range of 10 Hz to 500 MHz. It has an amplitude/phase accuracy of ±0.5 dB/±0.3 degrees for network measurements and greater than 70 dB dynamic range for spectrum measurements. Direct impedance measurements can be obtained using an optional impedance test set.

The instrument displays data in rectangular, tabular, Smith and polar display formats. It has a built-in 3.5 inch flexible disk drive for storing and retrieving control settings, measurement data, program-points table (for sweeps) and user programs. The 4195 can be customized to perform specific tasks with user defined functions and programs to improve efficiency.

For network measurements, standard HP accessories or user selected reflection/transmission test sets can be used to perform gain, phase, group delay, and reflection coefficient measurements. Its four input ports can perform spectrum measurements, allowing for multiple connections to a device without moving a test probe.

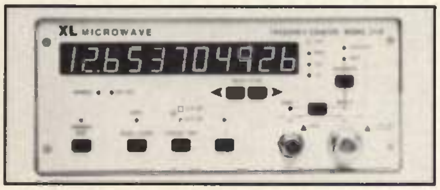


The HP 4195A network/spectrum analyzer is $23,000. A high stability, frequency-reference oscillator option for improved frequency-source stability is $850.

Also from HP is the HP 8904A multifunction synthesizer generates complex signals from a single chip. It can be used as a function generator,a stimulus for audio circuits, or as a modulation source for VOR, ILS, FM stereo and communications signalling. **Hewlett-Packard Company, Palo Alto, CA. INFO/CARD #174.**

10 Hz to 12.4 GHz Counter XL Microwave

The Model 3120 is a microwave counter that has a range of 10 Hz to 12.4 GHz with an acquisition time of less than 50 ms. Sensitivity is measured at −30 dBm. The instrument runs on either AC or DC and the price is $3300. **XL Microwave, Oakland, CA. INFO/CARD #173.**



DMOS Amplifier Linear Technology

LTI introduces the LB951 DMOS monolithic silicon RF amplifier. The specifications include a bandwidth of 10-200 MHz and 3rd order intercept of +34 dBm at 100 MHz. VSWR is less than 2:1 and S_{12} is less than −30 dB. The devices are packaged in either TO5 or plastic stripline packaging. **Linear Technology, Inc., Burlington, Ontario, Canada. Please circle INFO/CARD #172.**

900 MHz Power Module Motorola Semiconductor

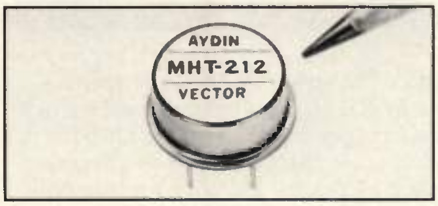
The MHW803 is a 2 W, 900 MHz power module for portable cellular radios. Also being introduced is the 2N6985 28 V, 125 W, 400 MHz and the 2N6986 28 V, 100 W, 500 MHz hermetic transistors. **Motorola Semiconductor Products Sector, Phoenix, AZ. INFO/CARD #171.**

Broadband Power Amplifier Instruments for Industry

The Model M5580 is a 10 kHz to 1.0 GHz amplifier with 75 W of output. The frequency range is broken down into two bands with the first being from 10 kHz to 500 MHz and the being from 500 MHz to 1.0 GHz. **Instruments for Industry, Inc., Ronkonkoma, NY. INFO/CARD #170.**

Hi-Rel RF Amplifier
Aydin Vector Division

The MHT-212 Series hybrid amplifier has a power output of 7 dBm with a bandwidth of 5 to 200 MHz. It is hermetically sealed in a 4-pin TO-8 package for direct insertion into microstrip boards. The typical gain is 22 dB and a reverse isolation of 25 dB is provided. It operates from a +15 VDC source at 25 mA.



The MHT-507 is essentially the same but with a gain of 14 dB, reverse isolation of 17 dB and requires a +5 VDC source at 7 mA. **Aydin Vector Division, Newtown, PA. INFO/CARD #169.**

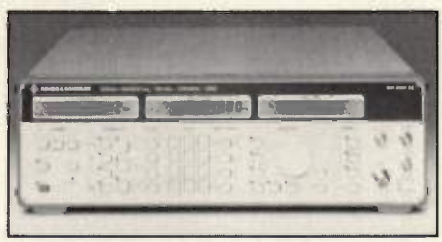
SMT Power Dividers
Synergy Microwave

Synergy introduces the SPD and SPD-C Series of hybrid surface mounted power dividers/combiners. The devices cover the .01 to 100 MHz, 1 to 500 MHz, 500 to 1500 MHz, 700 to 1000 MHz, and the 850 to 960 MHz range.

The SMF Series RF surface mounted filters are available in military and commercial versions. They are offered in low and high pass configurations at frequencies up to 500 MHz. **Synergy Microwave Corp., Paterson, NJ. INFO/CARD #168.**

Moderate Cost Signal Generator
Rohde & Schwarz-Polarad

The Model SMG has a constant resolution of 1 Hz over the 0.1 to 1000 MHz range. Broadband phase noise is less than −140 dBc at 100 MHz and spurious FM is below 2 Hz at 100 MHz. The instrument stores up to 40 panel setups for fast recall. IEEE 488 interface is standard. It is priced at $10,500. **Rohde & Schwarz-Polarad, Lanham, MD. INFO/CARD #167.**



MiCAD II
EEsof

MiCAD II is a computer aided design and layout program used to create drawings and generate artwork for microwave ICs. **EEsof, Inc., Westlake Village, CA. INFO/CARD #166.**

Color Monitor for Spectrum Analyzer
Marconi Instruments

Marconi introduces a color monitor option for its Model 2382 spectrum analyzer. The high resolution CRT has a 0.3 mm dot pitch. The spectrum analyzer covers 100 Hz to 400 MHz and features an integral tracking generator, automatic calibration, and built-in intelligence to solve measurement problems. **Marconi Instruments, Allendale, NJ. INFO/CARD #165.**

Digital Synthesizer/ Phase Generator
Sciteq Electronics

Sciteq's VDS-3 direct digital synthesizer covers DC to 3 MHz in one mHz steps. It is available with an internal phase control accessory that permits remote controlled phase rotation/modulation. With this option, phase can be rotated in steps of 0.36 degrees over ±360 degrees. BPSK and QPSK are fully supported. The VDS-3 is priced at $1995. Phase modulation and GPIB are $500 options. **Sciteq Electronics, Inc., San Diego, CA. Please circle INFO/CARD #164.**

Small Coax Terminations
EMC Technology

SMA coaxial terminations that dissipate up to 100 W and have dimensions of 0.5" × 1.0" are available from EMC Technology.

Maximum VSWR is measured at 1.5 from DC through 5 GHz. Impedance is 50 ohms nominal. In ten piece quantities, the 4180 is priced at $100 each. **EMC Technology, Inc., Cherry Hill, NJ. Please circle INFO/CARD #163.**

Matching Network Synthesis Program
Spefco Software

CIAO's synthesis option allows the design of lumped, distributed, mixed lumped distributed, lossless or lossy matching networks and filters between real or complex sources and loads over an arbitrary frequency range and with an arbitrary gain shape. **Spefco Software, Stony Brook, NY. INFO/CARD #162.**

Thick Film Resistors
JFW Industries

These resistors provide good RF properties up to 12 GHz. The use of thick film allows maximum film temperature up to 250 degrees C without significant resistance value change. **JFW Industries, Indianapolis, IN. INFO/CARD #161.**

SAW Spectrum Analyzer
Phonon Corporation

The AS 160-10-100 (C) MC is a SAW spectrum analyzer with a bandwidth of 10 MHz and resolution of 16 kHz. A high performance model with time sidelobe suppression greater than 40 dB is available. **Phonon Corporation, Simsbury, CT. INFO/CARD #160.**

RF/Microwave Test Equipment
Eaton Corp.

Eaton will be showing RF and microwave test and measurement equipment including EMI/EMS test instrumentation and systems, noise figure instrumentation, frequency synthesizers, spectrum analyzers, broadband/power amplifiers and AC instrumentation. **Eaton Corporation, Los Angeles, CA. Please circle INFO/CARD #159.**

RF Capacitors
Polyflon Company

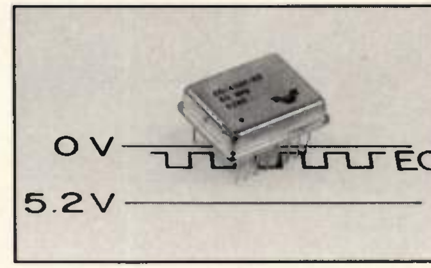
These capacitors feature compact and rugged PTFE construction with Q >5000. The capacitance range is 0.1 to 2000 pF. **Polyflon Company, New Rochelle, NY. INFO/CARD #158.**

Phase Locked Signal Source
Communication Techniques

Series-XPDRO is a dielectrically stabilized phase-locked signal source that uses a dual loop configuration to lock the output frequency to either a 5 or 10 MHz external reference. Power output is +15 dBm. **Communication Techniques, Inc., Whippany, NJ. INFO/CARD #156.**

DIP VCXO
Vectron Laboratories

The VCXO Series CO-434V is designed for phase locking applications. It provides ECL compatible output to any specified frequency from 8 MHz through 200 MHz. **Vectron Laboratories, Inc., Norwalk, CT. INFO/CARD #157.**



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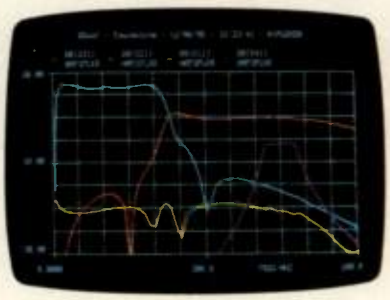


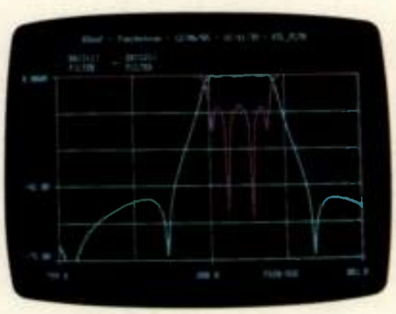
A

A Touchstone analysis of an active triplexer circuit with sampled output

B Insertion loss and return loss of crystal filter analyzed over its narrow pass band

C Schematic diagram of the active triplexer analyzed above

D A log axis can be used for displaying very broad frequency ranges. This feed back amplifier operates over nearly a decade bandwidth



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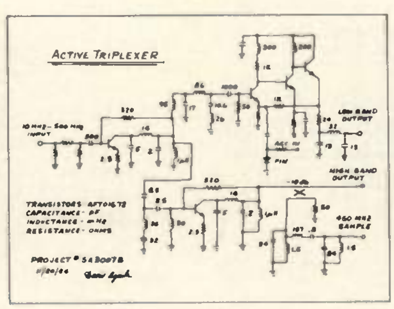

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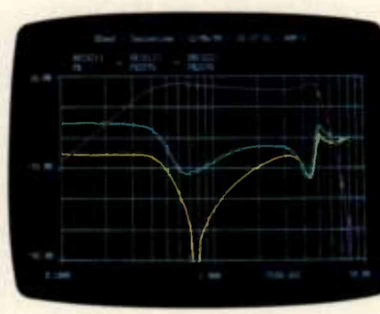

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Voltage Controlled Limiting Amplifier
Watkins-Johnson

The WJ-VCLA7 is a 50 to 500 MHz voltage controlled limiting amplifier in a TO-8 package. The output power limiting threshold can be varied by changing the control voltage. The WJ-VCLA17 is a similar device with a frequency range of 10 to 1000 MHz.

The MST201 is a single pole, double throw GaAs switch that covers DC to 5 GHz with insertion loss better than 1.8 dB and switching speed of 3 ns. It is hermetically sealed in a TO-5 package. **Watkins-Johnson Company, Palo Alto, CA. INFO/CARD #155.**

Hybrid TCXO
Reeves-Hoffman Division

Hybrid TCXOs are being introduced by Reeves-Hoffman. One has a center frequency of 10 MHz with stability of ±1 ppm and comes in a flatpack. The other has a center frequency of 4 to 100 MHz with the same stability and is packaged as a 24-pin DIP. **Reeves-Hoffman Division, Carlisle, PA. INFO/CARD #154.**

High Power Noise Source
Noise Com

Noise Com introduces a noise source that produces 50 W of white Gaussian noise from 10 to 175 MHz. The NC 9800

has a VSWR of 2:1 and is priced at $3900. **Noise Com, Inc., Hackensack, NJ. INFO/CARD #153.**

RF Power Meter
Bird Electronic Corp.

The Model 4420 digital directional RF power meter measures forward/reflected power in watts or dBm. It measures up to 1 GHz with power levels up to 1 kW. **Bird Electronic Corp., Cleveland, OH. Please circle INFO/CARD #152.**

Anti-aliasing Lowpass Filters
TTE

A line of subminiature anti-aliasing filters with a dynamic range greater than 80 dB has been announced by TTE. The fre-

quency range is from 10 kHz to 20 MHz. These filters cost $130 in single quantities. **TTE, Inc., Los Angeles, CA. Please circle INFO/CARD #151.**

Frequency Agile Filters
K & L Microwave

The Model 4HF-X225/X400-T/T frequency agile (hopping) filter utilizes 4 resonant sections and has a half dB bandwidth of 10 MHz with a typical 35 dB bandwidth of ±22 MHz. It measures 1.5" × 6" × 6".

Also from K & L is the Model 100 2×16 switch matrix that is suited for general purpose switching and actuation of external devices from DC to 50 MHz.

A 6 band selectable channel multiplexer with a frequency range of 850 to 1450 MHz will be show. The switching speed for the multiplexer is measured at 80 ns. **K & L Microwave, Inc, Salisbury, MD. INFO/CARD #150.**

Linear Hybrid Amplifier
TRW RF Devices

The Model CA2885 and CA2885H linear hybrid amplifier has a bandwidth of 40 MHz to 500 MHz with a gain of 18.5 dB. At the 1 dB compression point, the power output is 2 W. The bandwidth for the CA2838 and CA2838H is 100 kHz to 325 MHz and the output at the 1 dB compression is 900 mW.

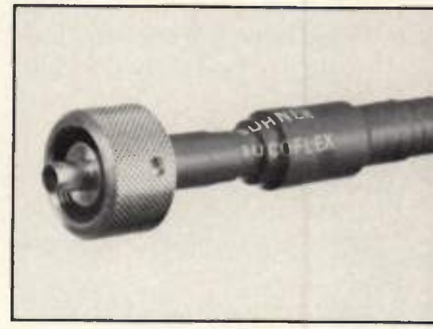
Also being introduced is a Class A UHF power amplifier (PAA-0810-40-50LAC) with 40 dB gain and 50 W output. The bandwidth ranges from 800 MHz to 1000 MHz

while the noise figure is measured at 8 dB.

RF amplifier modules with power outputs from 300 mW to 4 W and 14 dB to 36 dB gain will be unveiled. Five bandwidths ranging from 200 MHz to 990 MHz over a frequency range of 1 MHz to 1 GHz are available. **TRW RF Devices, Lawndale, CA. INFO/CARD #149.**

Microwave Cable Assembly Huber + Suhner

The Huber + Suhner Sucoflex is a cable assemble compatible to HP 8510 network analyzers. The assembly employs a flexible tube of stainless steel that protects it against pulling, crushing and abrasion. **Huber + Suhner, Inc., Woburn, MA. INFO/CARD #148.**



Oscillator with good Spectral Purity Frequency and Time Systems

The FTS 1000B oscillator has a SSB phase noise of −147 dBc at 100 Hz and −160 dBc at 10000 Hz. The standard FTS1000B at 5 MHz with 4 outputs is priced at $2250. **Frequency and Time Systems, Inc., Beverly, MA. Please circle INFO/CARD #147.**

DPDT Relay FL Jennings

A DPDT relay for applications to 3.2

GHz is being introduced. It requires a .100" grid space pattern and does not need spreader pads. Depending on quantity, the price ranges from $63 to $20. **FL Jennings, San Jose, CA. Please circle INFO/CARD #219.**

Portable Spectrum Analyzer
Avcom of Virginia

The PSA-35AD portable spectrum analyzer with wideband FM audio demodulator is designed for analysis of wideband satellite communication signals. It offers frequency coverage of 10 to 1750 MHz and 3.7 to 4.2 GHz. It has a standard frequency band calibrated from 1250 to 1750

MHz to cover European BCD frequencies. The price with FM demodulator is $2415. Without the FM demodulator it is priced at $1965. **Avcom of Virginia, Inc., Richmond, Va. INFO/CARD #217.**

Microwave Harmonica
Compact Software

Harmonica performs analysis and optimization of non-linear analog circuits under single or multitone excitations.

Super Spice is a non-linear program based on 2G3 SPICE that provides a large library of components. In addition to the standard elements, GaAs FET modelling is possible.

TRL Design Kit includes the synthesis and analysis of transmission line parameters for stripline, microstrip, coax, rectangular and ridge waveguide.

Coupler Design Kit provides electrical and dimensional synthesis of Lange couplers, edge-coupled stripline and microstrip couplers, and multihole topwall couplers in rectangular waveguide.

Sonata is a CAD package for synthesis and design of microwave oscillators. The program determines the best solution and provides, among other things, a component list. **Compact Software, Inc., Paterson, NJ. INFO/CARD #211.**

SAW Oscillators
RF Monolithics

A companion receiver for the previously announced microtransmitter is being unveiled. Also to be shown will be SAW oscillators for applications up to 1.5 GHz. **RF Monolithics, Inc, Dallas, TX. Please circle INFO/CARD #207.**

Broadband Amplifiers
Amplifier Research

Models 25W1000M7 and 10W1000M7 cover from 100 to 1000 MHz and deliver 25 W and 10 W respectively. Both models

offer instantly available bandwidth for sweep testing. **Amplifier Research, Souderton, PA. INFO/CARD #142.**

UHF ISOFETs
Acrian

The UMIL20FT and UMIL40FT are UHF ISOFETs for use in the 225 to 400 MHz range. They exhibit 20 and 40 W of output respectively. **Acrian, Inc., San Jose, CA. INFO/CARD #141.**

GaAs FETs
California Eastern Labs

The NE345L-10B and NE345L-20B are GaAs FETs with the 1 dB compression point output at 40 dBm and 43 dBm, respectively. The 10B has a gain of 9 dB at 2.3 GHz while the 20B has a gain of 10 dB at 1.6 GHz.

A hetero junction GaAs FET (NE20383A) with a 13 dBm power output at the 1 dB compression point is being unveiled. The gain is 10 dB and the noise figure is 1.25 dB. **California Eastern Laboratories, Inc., Santa Clara, CA. INFO/CARD #140.**

Beam Lead GaAs Tuning Varactors
M/A-COM

These beam lead GaAs hyperabrupt tuning varactors feature constant gamma. The high Q values and the elimination of package parasitics make them useful for broadband VCO's that require linear frequency tuning.

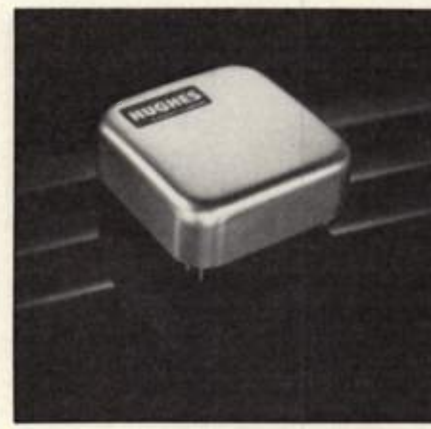
The MA4GPXXX family of GaAs beam lead PIN diodes offer improved microwave characteristics for switching applications. The high carrier mobility of GaAs results in low series resistance and fast switching. They may be switched directly using TTL signals.

The M/A 4GMJ01 voltage variable absorptive attenuator is optimized for operation in the DC to 2 GHz range. The insertion loss is typically 1.3 dB with the attenuation range approximately 20 dB.

Also being introduced it the M/A 4GM203 and 4GM204 GaAs MMIC switches. The units are configured as SP3T and SP4T, respectively. Depletion mode MESFETs are used to insure flatness and linearity of the switches. **M/A-COM, Inc., Burlington, MA. INFO/CARD #139.**

uP Controlled Crystal Oscillator Hughes Aircraft Co.

By combining advancements in crystal, CMOS and microprocessor technology, Hughes has created the microprocessor controlled crystal oscillator (MCXO). With



this it is possible to develop a TCXO with the thermal characteristics similar to low stability OCXOs, but requiring 1/10th the power. **Hughes Aircraft Co., Newport Beach, CA. INFO/CARD #138.**

Thick Film ICs KDI Electronics

KDI has expanded its thick film microwave integrated circuit capability to include sub-assemblies. It ranges from elementary switches to PIN diode switches and monolithic drivers. **KDI Electronics, Inc., Whippany, NJ. Please circle INFO/CARD #137.**

35 MHz NCMO Digital RF Solutions

The DRFS-3250 is a number controlled modulated oscillator with 12-bit phase output resolution. It has a clock frequency to 35 MHz with sinusoid synthesis from DC to 17.5 MHz. Phase modulation resolution exceeds 0.1 degrees. Input interface circuitry is included to simplify application in three basic tuning modes: 24-bit direct parallel, 8-bit bus, and quadrature clocked serial. The device uses digital techniques to provide a low noise signal source which simultaneously features high stability, fine resolution and rapid switching. **Digital RF Solutions Corp., Santa Clara, CA. INFO/CARD #178.**

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The Sub-System Design Group at Merrimac has established an extensive library of the common types of microwave stripline devices. All the critical design parameters of each component are stored on computer disk, immediately available for recall by the CAD-CAM for direct integration into each new sub-system design.

The library of component parts includes careful designs of the basics, such as idealized SMA to stripline interfaces and 50 ohm transmission paths as well as the conventional component. Each is stored in several optimized versions to cover various frequency bands and desired system parameters. Designs on file include:

Two-way Power Dividers
True Three-way Power Dividers
True Four-way Power Dividers
True Five-way Power Dividers
6, 10 and 20 dB Couplers
Amplitude Equalizers
Quadrature Hybrids
Phase Shifters

In the Microwave Combiner/Equalizer pictured above, a total of 14 stripline components were integrated into a single assembly. Had the assembly used individual components, it would have comprised the following:

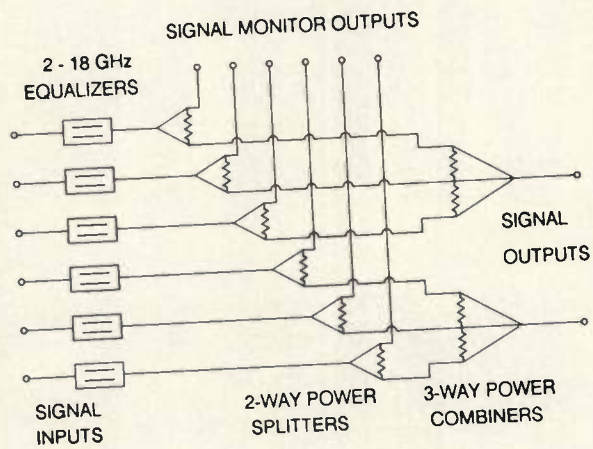
Qty. 6 - Two-way Power Dividers
Qty. 2 - Three-way Power Combiners
Qty. 6 - Amplitude Equalizers
Qty. 12 - Phase Matched Cables
Qty. 1 - Mounting Plate, etc.

Such an assembly would likely weigh 2 lbs. (1.8 kg) or more, occupy much space and present the user with a number of difficult mounting and reliability problems.

For a compact design Merrimac sandwiched together two identical halves of the required circuit into a single assembly weighing just 440 grams, and measuring a compact 5.0" x 3.9" x 0.65" (120 x 94 x 16 mm), plus connectors.

Electrically the benefits of this vertical integration included reduced insertion loss, predictable high isolation between ports, and inherently high reliability. In this example, the low insertion loss was masked by the 10 dB slope equalizers, however, crosstalk isolation between channels was greater than 20 dB. Input VSWR was maintained at better than 2.0:1 in spite of the effects of the slope equalizers. All specifications were maintained over the required temperature range of -40 to +90°C, as well as other uninhabited airborne environmental specifications.

For further details on this and other stripline sub-systems contact MERRIMAC INDUSTRIES, INC.



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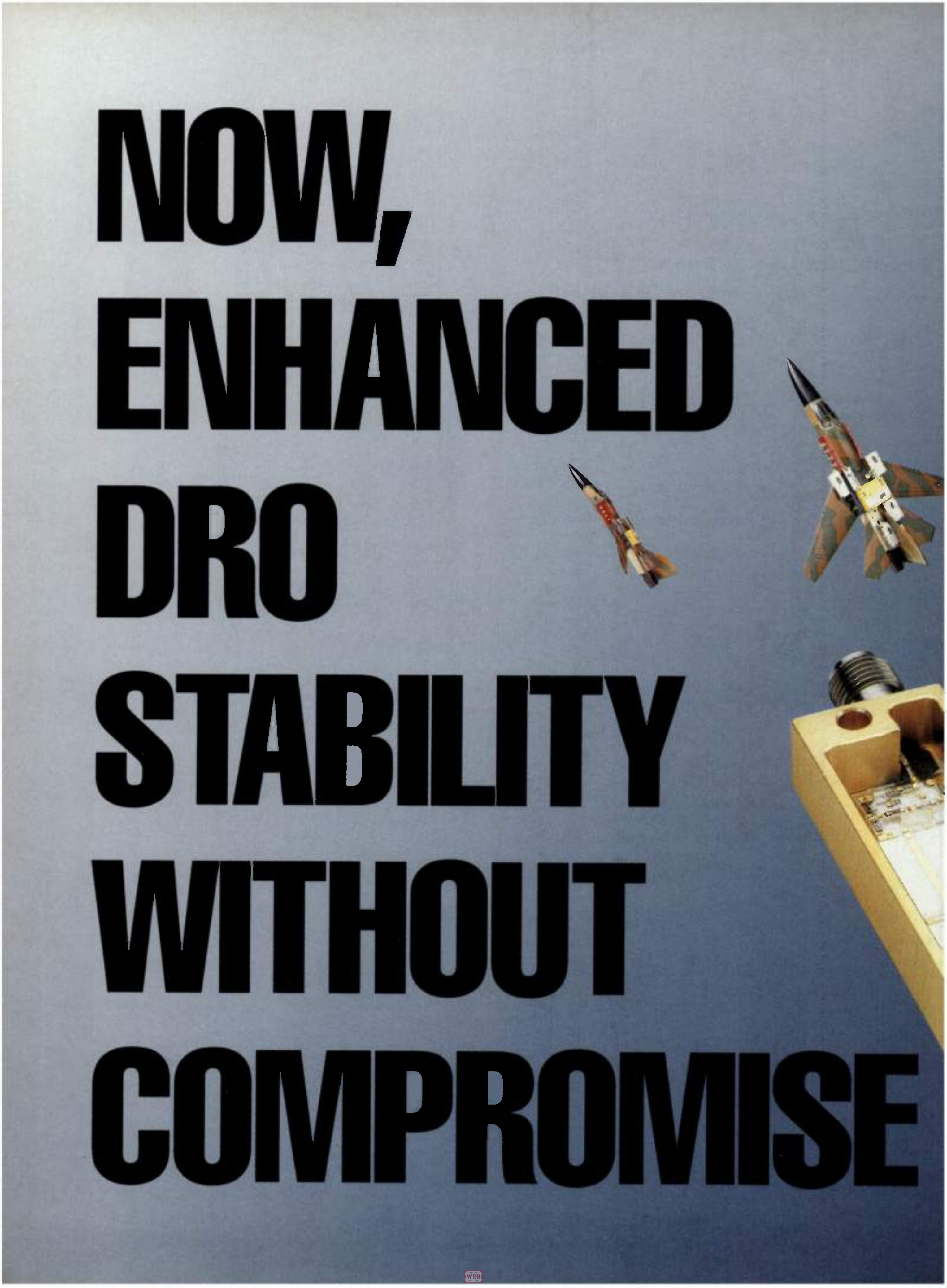
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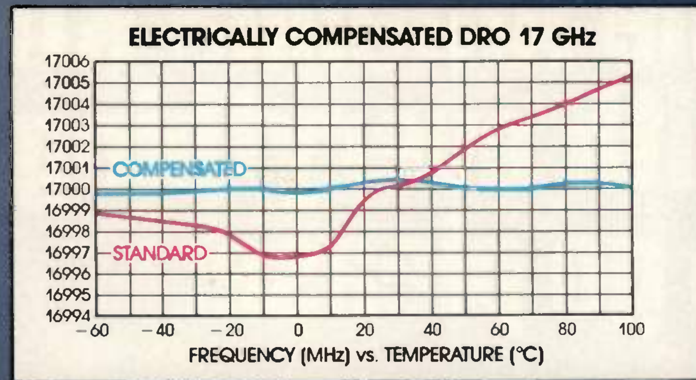

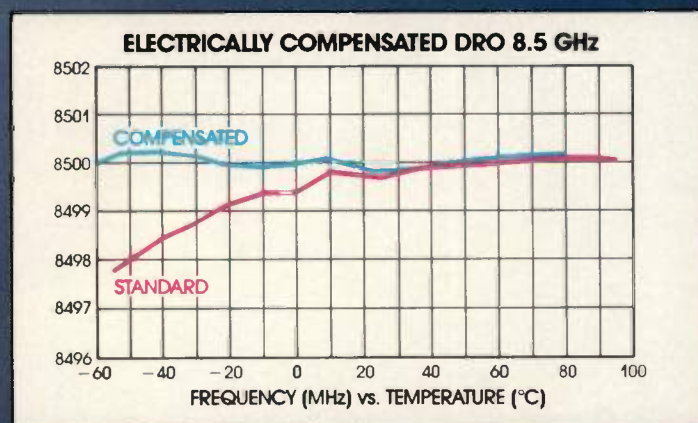

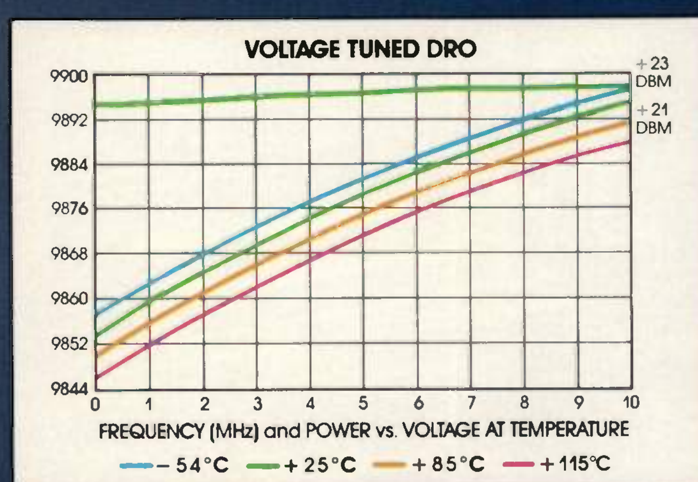

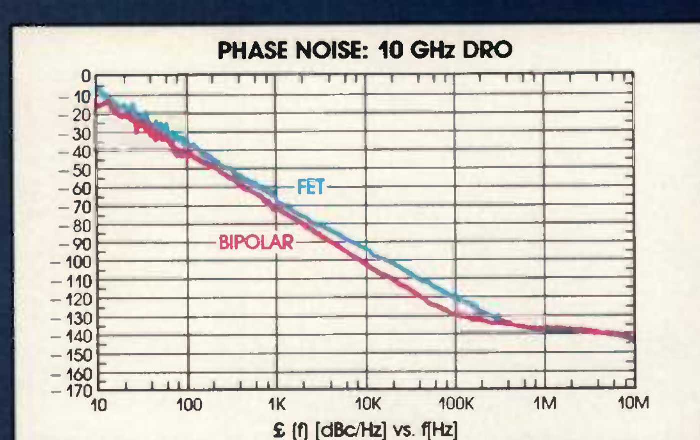

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INFO/CARD 56

20-Year Evaluation of Shielding Tape

Facility Construction Offers Rare Opportunity to Measure Long-Term Effects.

By Richard H. Jackson
Electrical Products Division, 3M

In the shielding tape business, it is rare to have an opportunity to examine permanently installed tape products after more than a few years. In July 1986, the Technical Service Section of 3M's Electrical Products Division had an opportunity to examine tape that had been installed nearly 20 years previously and had remained virtually untouched since its installation.

The structure to be refurbished was the RF building at the Naval Weapons Center, built in 1967 at China Lake, California. The building originally was shielded and subsequently refurbished. During the refurbishing, 3M technical service personnel were permitted to take comprehensive measurements on the foil tape originally used to shield the building, in order to compare the results with tests of new equivalent products.

Most of the RF building (Building 31440) at China Lake consisted of a large RF anechoic room. The remainder housed offices and test equipment. The interior of the finished anechoic chamber was 110 feet long, 40 feet wide and 40 feet high. It consisted of a double-walled, shielded metal structure that was lined on the interior with RF absorbing foam blocks (treated plastic foam wedges and cones).

The double-wall construction consisted of an exterior metal wall that was not sealed to prevent RF leakage, but nevertheless provided some shielding. The inner wall was made of galvanized steel on which all seams and joints were taped with Scotch brand copper foil shielding tape No. 1181. This tape consists of a copper foil backing with a conductive adhesive system.

The inner wall consisted of double-box sections of 22-gauge galvanized steel. Each panel was approximately one foot wide and 10 feet tall. A half-inch strip of 1181 tape covered each seam between these panels to the full height of the room — some 50 feet or so overall. The gap between the panels, which the half-inch tape bridged, was approximately one-eighth inch. A 1-inch tape was used between the panels at the horizontal seams.

Test	China Lake Results	Typical Values
1. Shear CF	<1.75 mm/hr @10.3 N/cm	<1.00 mm/hr 10.3 N/cm
2. Peel CF 180 deg	<12 mm/min @6.2 N/cm	<250 mm/min @6.2 N/cm
3. Peel CV 90 deg	8.2 N/cm (avg) @30 cm/min	4.8 N/cm (avg) @30 cm/min
4. Peel CV 180 deg	10.9 N/cm (avg) @30 cm/min	4.4 /cm (avg) @30 cm/min
5. Resis	20 to 150 ohms/sq cm	10 to 100 ohms/sq cm
Test temperature	@105 degrees F	@73 degrees F

Table 1. Results Tabulated with Product Specifications.

The concrete floor was covered with heavy gauge steel panels, each 4 by 8 feet in size and seamed with 1-inch tape. The RF-absorbing foam covered the floor, walls and ceiling. Throughout the room, the foam was attached over the galvanized steel and copper tape with industrial-grade, rubber-base adhesive.

Most of the anechoic RF test facility had not been altered since its construction in 1967. The areas in which the foil tape was examined were covered with anechoic foam material until the 1986 refurbishing. During the building's 19-year existence, it had been exposed to the typically hot and arid conditions of the Mojave Desert.

Although the RF building interior was air conditioned, the surface on which the tape had been applied had been exposed to significant temperature extremes and cycles. This was due in part to double-wall construction without air circulation between walls, and the outer wall's direct exposure to the desert sun. The insulating nature of the RF absorbing foam contributed to this effect by isolating the interior wall from the cooling effects of the interior air conditioning.

Just as the taped inner wall was exposed to the heat of the sun in the day, it also was exposed to rapid cooling typical of the high desert — including some below freezing temperatures in winter. The high temperature to which the inner wall was exposed was in the range of 120 degrees F to 150 degrees F — the same temperature as the dead air space between the walls. This daily high range occurred year-round, while the lows ranged from about 60 degrees F at night in the summer to as low as 25 degrees F in the winter.

The room was exposed to very dry conditions through most of its history, with one exception: a few years ago it was accidentally exposed to a large amount of water as a result of frozen and burst pipes in the fire protection system. This water exposure was significant and lengthy since the foam held great quantities of water like a sponge. The moisture produced corrosion on walls and the floor of the building.

The tests performed on the tape at the site included:

- static shear (constant force)
- 180-degree peel adhesion (constant

force)

- 90-degree peel adhesion (constant velocity)
- 180-degree peel adhesion (constant velocity)
- Electrical resistance

These tests were chosen to gain insight into the performance characteristics of the old tape and to compare it to new products, and are not normally run on all tapes manufactured. Only tests numbered 4 and 5 are part of the normal screen tests done on each lot of tape. Test results in the table below were obtained for new tape for comparison purposes specifically for this analysis.

In areas that obviously had been dry throughout the building's history, construction materials appeared corrosion-free and almost new, with the exception of some areas where there were hand and finger prints. These probably were left from the time of construction.

The areas exposed to water over a prolonged period of time were quite corroded. The copper surfaces were dark-green and almost black in some areas. On some parts of the floor, both copper tape and steel were completely missing. This corrosion of steel panels on the floor may have been accelerated by the chemical action of the concrete slab below.

The surfaces in contact with the adhesive, the interface between galvanized steel and copper tape appeared to have been preserved and protected by the action of the adhesive, with the exception of areas in which either the tape or the steel were corroded away entirely.

In the undamaged areas, the tape's adhesive was still very tacky, and appeared to have all the properties of new tape. In some areas, when the tape was peeled from the host surface, the adhesive stayed on the steel and did not come off with the copper backing. This is not an unusual circumstance, since adhesion to copper and steel is about the same after aging.

The tape and its adhesive performed very well over the life of the building. Considerably more local corrosion might have been expected between the copper and the zinc than observed. In an installation of this type in any other environment, it might be wise to use the tin-alloy coated, embossed product (Scotch brand 99 tape) to avoid galvanic corrosion.

The protective characteristics of the adhesive were clearly demonstrated by the cleanliness observed on the surfaces covered by the adhesive. There were no signs of corrosion or other chemical interaction between the metal surfaces and the adhesive.

The test results obtained show clearly that the adhesion of the tape does not degrade with time, but rather increases significantly. This is particularly true for the dead-load peel.

The measurement of electrical resistance showed that the electrical contact is maintained through the adhesive. Results obtained are typical of short-term installations of this type of tape on this substrate.

More information on this article can be obtained by calling (800) 233-3636 or (612) 733-5699. rf

About the Author

Richard H. Jackson is Electrical Engineering specialist at the Electrical Products Division, 3M Center, Bldg. 260-5B-04, St. Paul, MN 55144. He has 22 years of experience at 3M.

PLL Primer — Part IV

Description of a Loop Integrator Circuit.

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

This is the fourth part in a series on PLL primers that began in 1983. Part I, II and III (Ref. 1, 2 and 3) dealt with some of the overall design aspects of modern PLL circuits. This part describes the characteristics of the loop integrator circuit. Some of the design aspects are reviewed and its effect on the slew rate of the PLL are analyzed.

Most of the modern PLL circuits are of the type 2, employing an active integrator as the filter after the phase comparator. Theoretically, an ideal integrator has infinite gain at DC and a finite gain decreasing with frequency, which is determined by its time constant(s). Conventionally, the PLL order is determined by the integrator order. This is an approximation, neglecting the additional loop poles. Figure 1 shows the first, second and third order loop integrators. The first order is seldom used and the second does not usually result in a second order PLL since the VTO provides the additional pole. The third order PLL is usually the lowest realizable PLL order.

The integrator characteristics are determined by the integrator time constants and the characteristics of the active device. The active device, usually an operational amplifier, is assumed to be ideal, i.e. it has infinite DC gain and infinite response. In most narrow band loops this assumption is valid, but for critical applications, the finite gain and response have to be taken into consideration.

These operational amplifier imperfections cause problems when used with digital phase comparators. These comparators provide a correction output in the form of very short pulses of either polarity. These pulses, especially when ECL comparators are used, have very steep rise and fall times causing the operational amplifier to integrate them inaccurately. The integrator configuration is changed to its alternative equivalent form, shown in Figure 2. The input pulses are then partially integrated by the input RC time constant. These pulses, with considerably decreased rise and fall times, can then be handled by the active integrator.

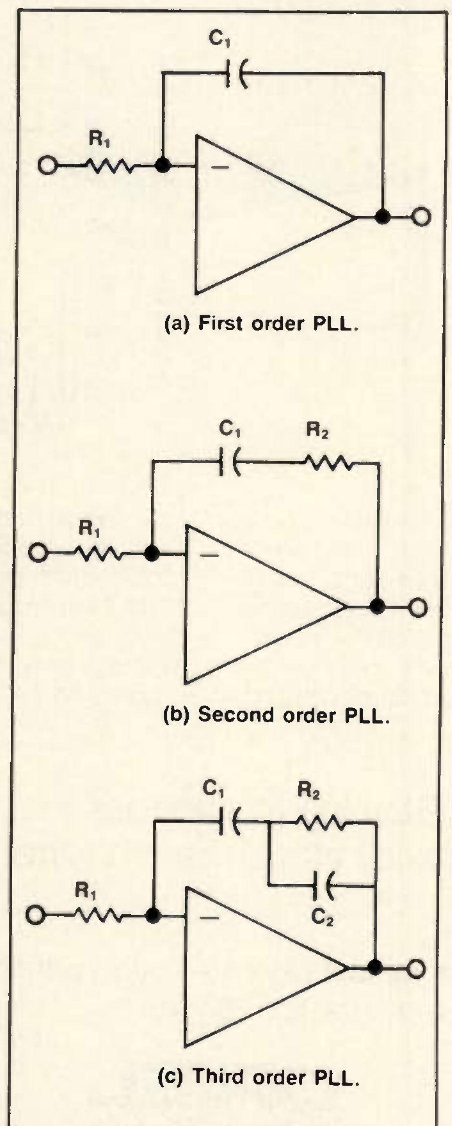


Figure 1. PLL active integrators.

The operational amplifier finite characteristics also affect its frequency response. A typical operational amplifier may have a DC gain, A_o, of about 100,000 and its first pole, F_o, at about 100 Hz. The overall response, including these two imperfections (for a third order PLL integrator) is:

$$F(S) = \frac{S \cdot A_o (T_2 + T_3) + A_o}{S^3 \cdot B + S^2 \cdot C + S \cdot D + 1} \quad (1)$$

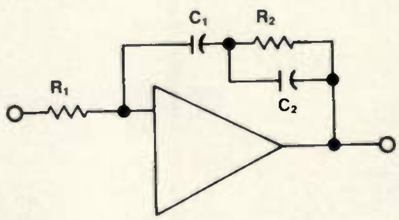
where

$T_1 = R_1 \cdot C_1$
$T_2 = R_2 \cdot C_1$
$T_3 = R_2 \cdot C_2$
$T_o = \frac{1}{(2 \cdot PI \cdot F_o)}$
$B = T_o \cdot T_1 \cdot T_3$
$C = A_o \cdot T_1 \cdot T_3 + T_1 \cdot T_3 - T_o (T_1 + T_2 + T_3)$

$D = A_o \cdot T_1 + T_o + T_1 + T_2 + T_3$

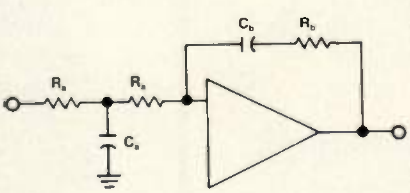
In terms of frequency, Equation 1 becomes:

$$F(\omega) = \frac{j(\omega \cdot A_o(T_2+T_3))+A_o}{j(\omega \cdot D - \omega^3 \cdot B) + (1 - \omega^2 \cdot C)} \quad (2)$$



$$T_1 = R_1 \times C_1$$
$$T_2 = R_2 \times C_1$$
$$T_3 = R_2 \times C_2$$
$$F(s) = \frac{1 + s(T_2 + T_3)}{sT_1(1 + sT_3)}$$

(a) Conventional.



$$T_a = R_a \times C_a$$
$$T_b = R_a \times C_b$$
$$T_c = R_b \times C_b$$
$$F(s) = \frac{1 + sT_c}{2\,s\,T_b(1 + sT_a/2)}$$

$$T_c = T_2 + T_3$$
$$T_b = (T_1)/2$$
$$T_a = 2 \times T_3$$

(b) Modified.

Figure 2. Integrator configuration.

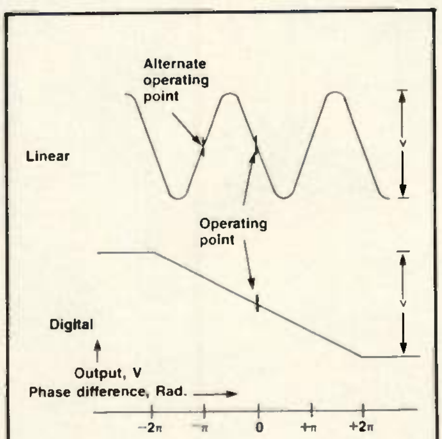


Figure 3. Phase comparator characteristics.

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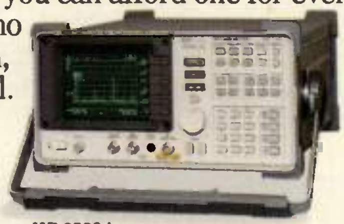

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Newton refracts light through a glass prism, circa 1672. "The Bettmann Archive."

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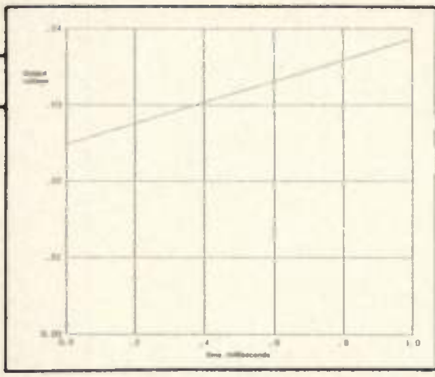

Figure 4. Second order PLL integrator time response.

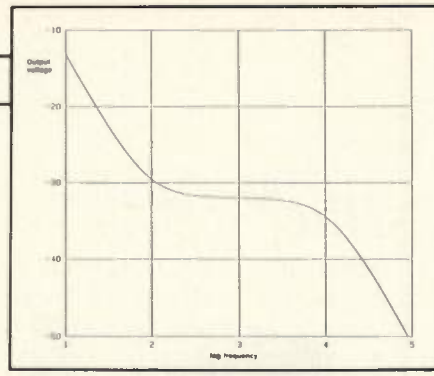

Figure 5. Third order PLL integrator frequency response.

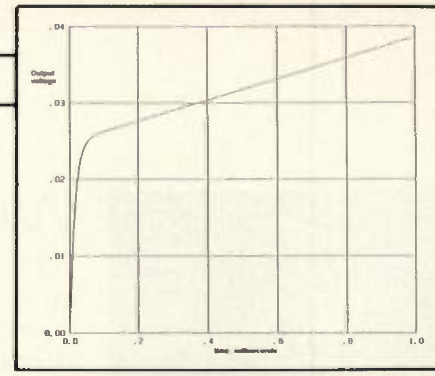

Figure 6. Third order PLL integrator time response.

The speed with which a PLL can sweep over a certain frequency range is mainly determined by the integrator. Thus, it is important to understand this mechanism when fast frequency changes are required. To facilitate discussion, let's assume an ideal digital phase/frequency comparator, as shown in Figure 3. This ideal comparator has a voltage output proportional to phase difference during phase acquisition, and a constant maximum (negative or positive) output voltage during frequency acquisition. First the second order integrator will be considered. The time constants are:

$T_1 = 0.0728$
$T_2 = 0.00181$

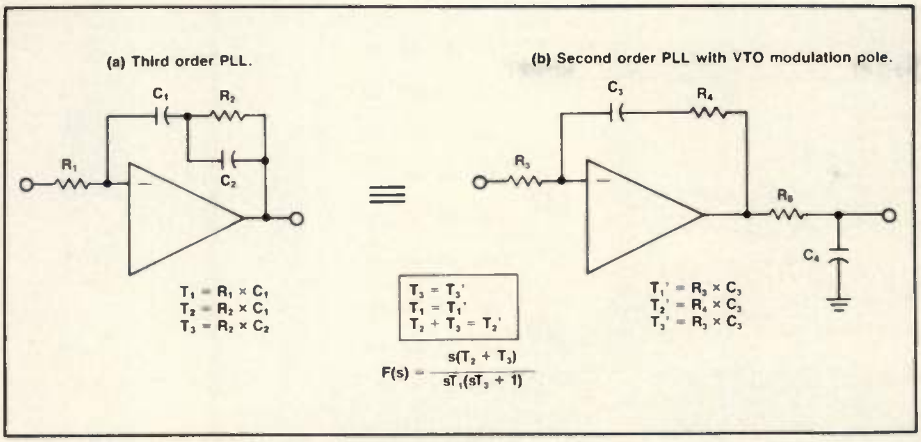

Figure 7. Equivalent integrator circuits.

The integrator output versus time, for a constant input voltage, V, is

$$V_o = V\,(t/T_1 + T_2/T_1) \qquad (3)$$

where t = time after voltage, V, was applied.

Plotting the above for a one millisecond period we obtain the curve shown in Figure 4. Adding $T_3 = 0.0000139$, which makes it a third order PLL integrator, and including the frequency response shown in Figure 5, equation 3 becomes

$$V_o = (t/T_1 + (T_2 - T_1\,(1 - e^{-t/T_3}))) \qquad (4)$$

The plotted data is shown in Figure 6. Thus, for small changes in frequency (small change in integrator output, or VTO input voltage), it seems that the second order integrator has an advantage since its output voltage changes abruptly while

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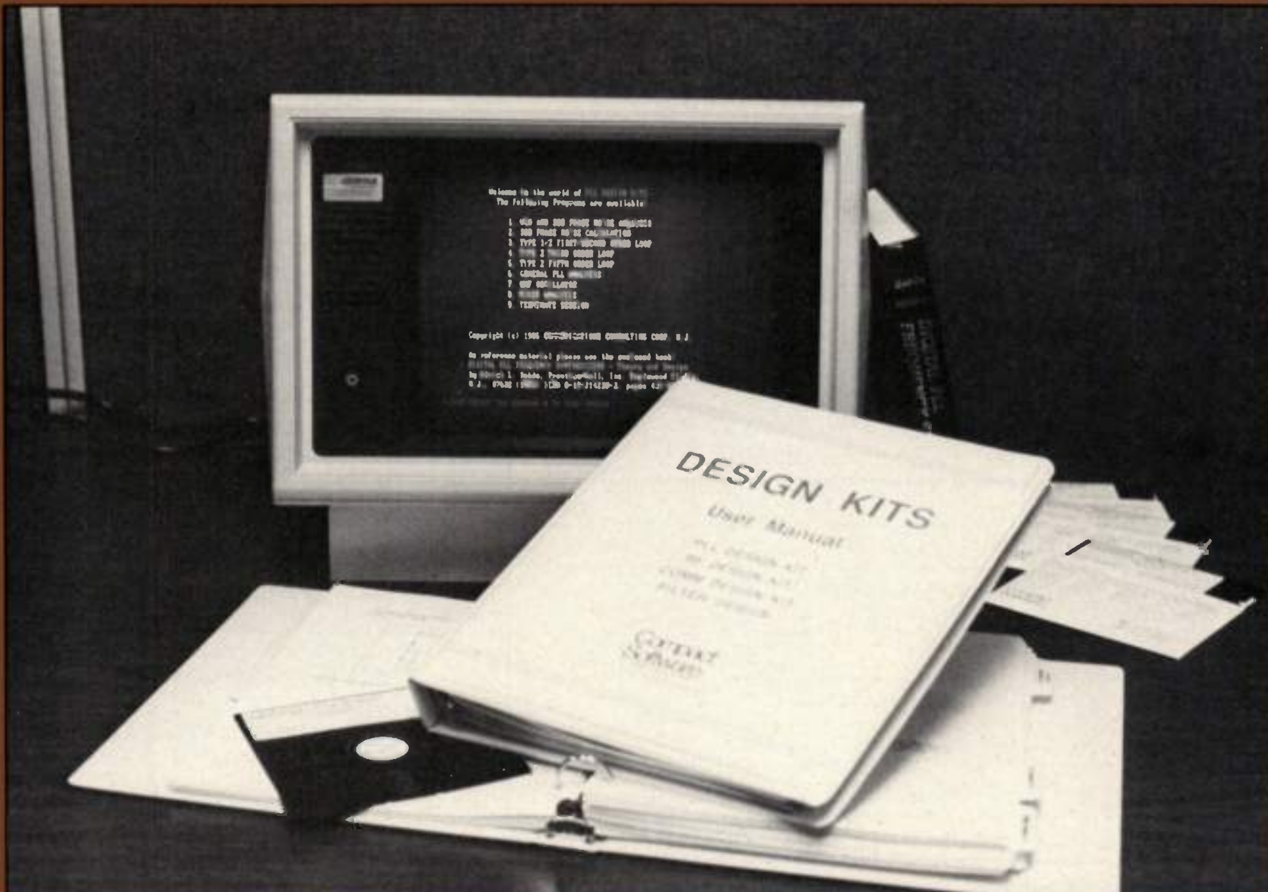

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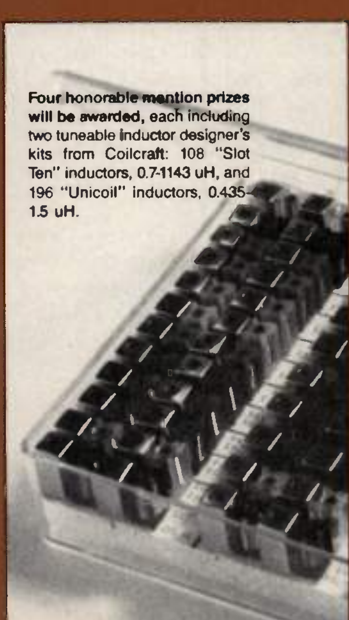

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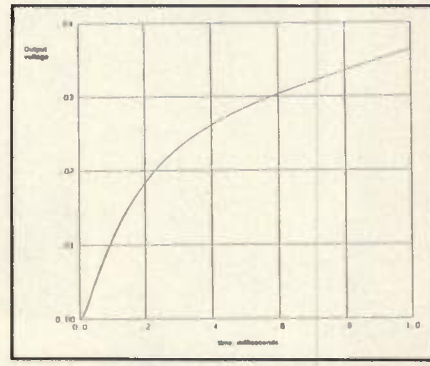

Figure 8. Third order PLL integrator with VTO pole time response.

the third order integrator has some initial delay. However, the second order overall response cannot be obtained in a practical circuit because of the VTO modulation input pole.

It can be shown that a second order integrator followed by a VTO lowpass modulation input characteristic is equivalent to a third order integrator. This is seen in Figure 7. The obvious question then is: how does the VTO input affect the third order integrator response to a step input function? This is shown in Figure 8 for a VTO modulation bandwidth of 1000 Hz. To provide a complete picture, the following additional circuit imperfections were added:

Integrator op-amp open loop gain — 100,000
Integrator op-amp first pole — 10 Hz

Unfortunately, this calculation is quite involved, when all of the above factors are considered.

From the above discussion it is apparent that, to optimize the PLL switching speed for small frequency changes, all the factors have to be carefully considered. The usual approximations may lead to erroneous conclusions. rf

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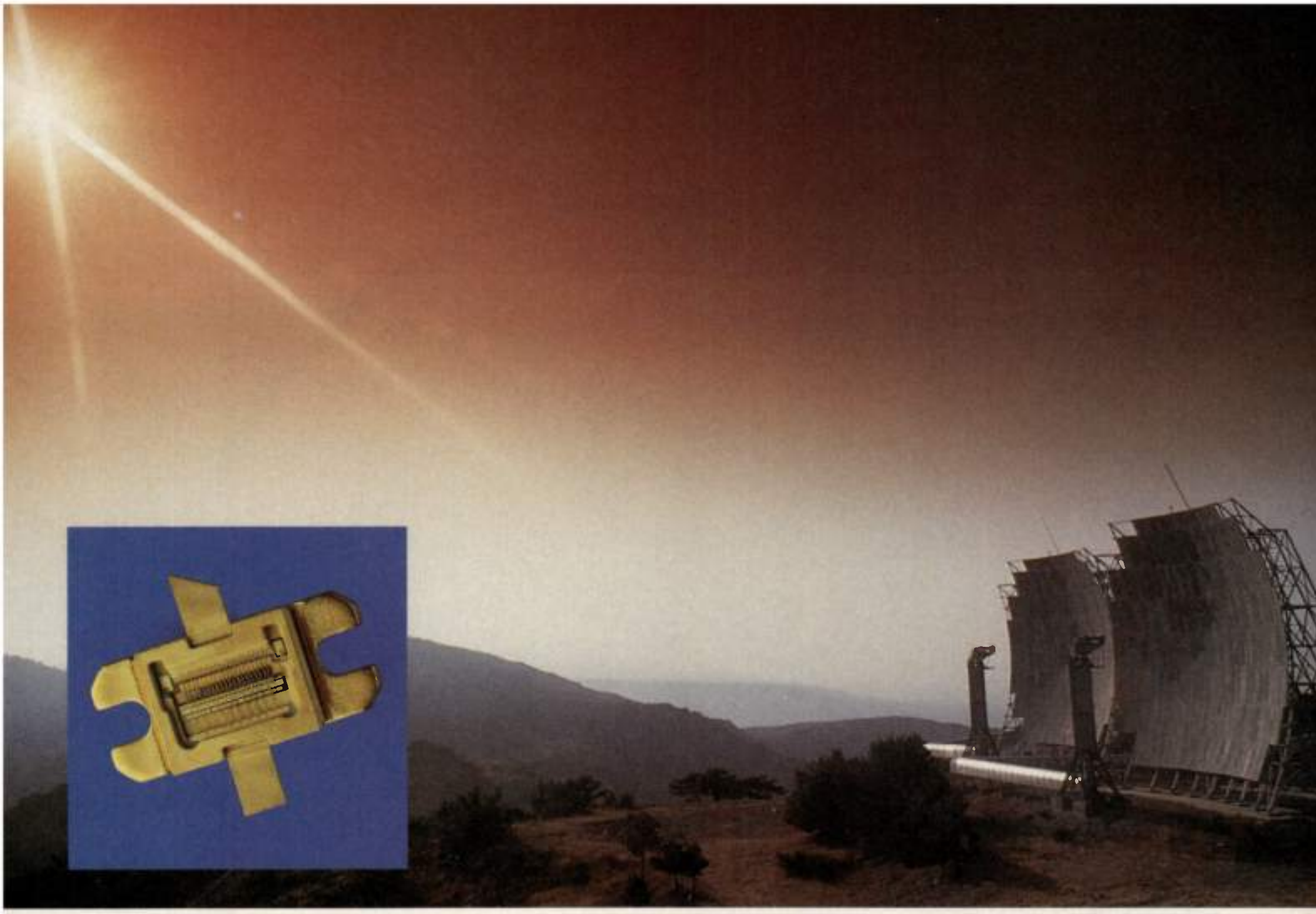
About the Author

Andrzej B. Przedpelski is vice-president, development of A.R.F. Products, Inc., 2559 75th St., Boulder, Colorado 80301. He serves as consulting editor to *RF Design*.



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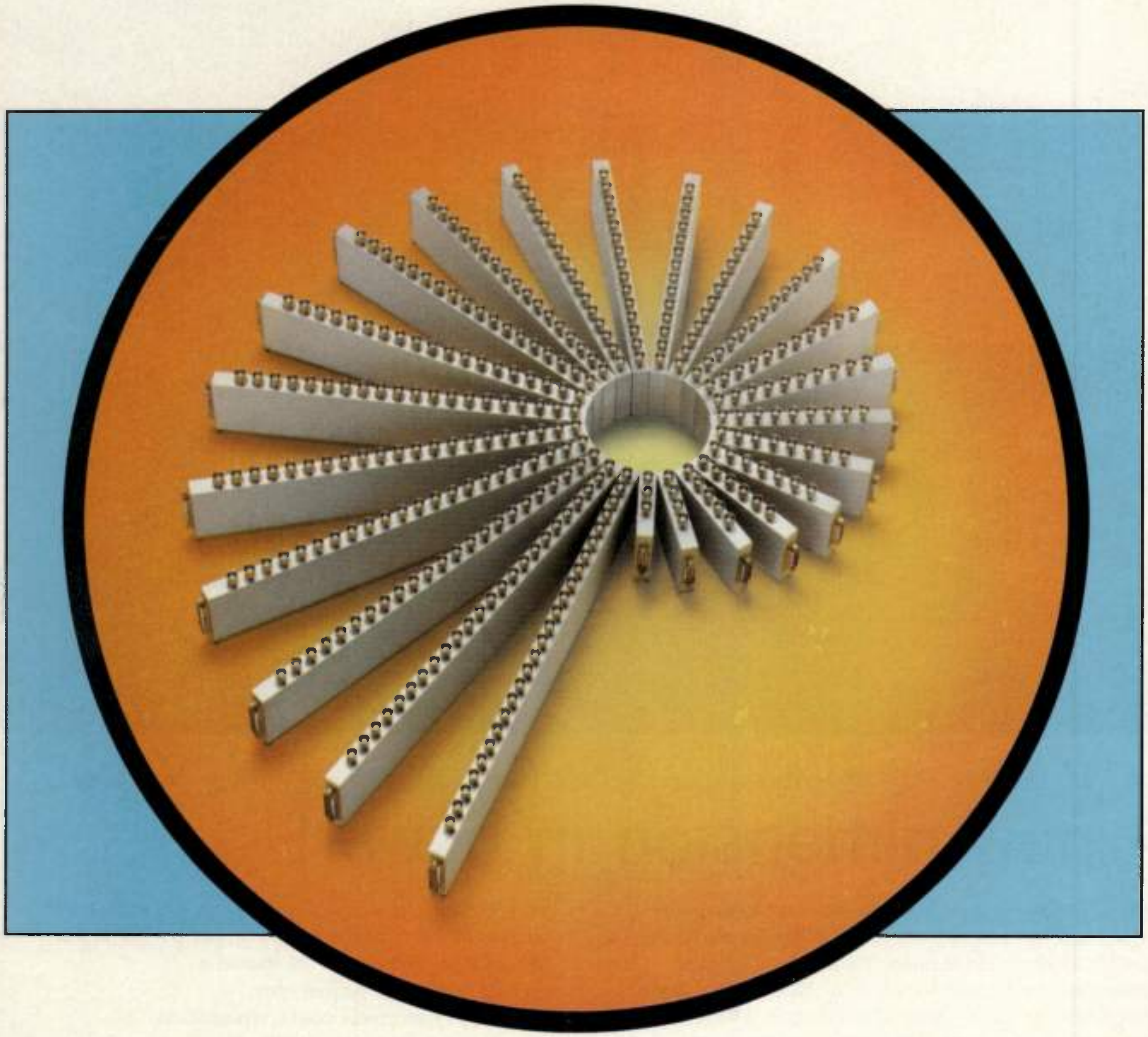

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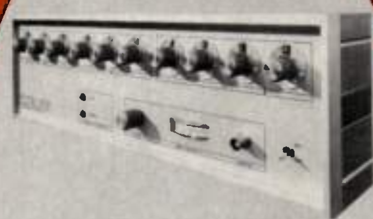

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Broadband Microwave Transistor Amplifier Design Using S-Parameters

By Jackie L. Hughes
Spectradyne

Scattering parameters are becoming widely used because they are much easier to measure and work with than Y-parameters. They are easy to understand, convenient, and provide a wealth of information at a glance (2). This article utilizes S-parameters in designing a broadband microwave transistor amplifier.

The design procedure for a microwave transistor amplifier using S-parameters is presented. The following design goals are specified:

Center Frequency:	1.5 GHz
Bandwidth:	1-2 GHz
Noise Figure:	less than 3.5 dB
Gain:	10 dB (small-signal)

The transistor used will be a bipolar junction transistor (BJT). The amplifier will consist of a single stage, operating into a 50 Ω load and driven by a 50 Ω source.

After choosing an appropriate transistor, a suitable DC biasing circuit is designed at a specific collector current and collector-emitter voltage as recommended by the vendor. The Hewlett-Packard HXTR-3101 general purpose transistor is selected due to its excellent gain characteristics and low noise figure. The HXTR-3101 is also unconditionally stable in the 1-2 GHz band.

The bias conditions recommend by HP for the HXTR-3101 for minimum noise figure shown on the data sheet are:

$$V_{CE} = 10 \text{ V} \quad (1)$$
$$I_C = 10 \text{ mA} \quad (2)$$

Microwave transistor amplifier design requires biasing the transistor into the active region of performance(8). At low frequencies, emitter resistor stabilization with negative current feedback is used for DC stability. Most microwave circuit designs for best gain or lowest noise figure require that the emitter lead be DC grounded as close to the package as possible so that the emitter series feedback is kept to an absolute minimum (8).

Two bias circuits recommended by Hewlett-Packard are shown in Figures 1A and 1B(8). The circuits in both figures find widespread usage as bias networks. The voltage feeback circuit (Figure 1A) uses fewer components and is almost as temperature stable in Figure 1B. For simplicity, the circuit of Figure 1A will be chosen for use since the circuit of Figure 1B is only slightly more stable with temperature(8).

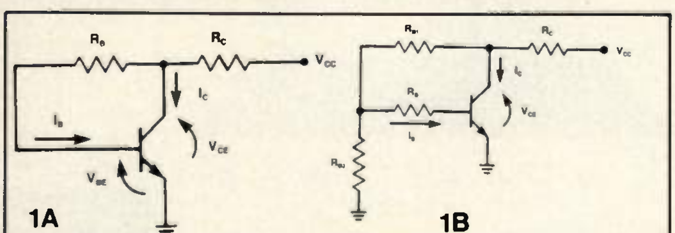

Figure 1A. Voltage feedback.
Figure 1B. Voltage feedback with constant I_B.

A conveninent value for V_{CC} in the bias circuit is 20 V (regulated from a standard 24 V power supply).

Then, $V_{CE} = 10$ V
$I_C = 10$ mA
Assume $V_{BE} = .7$ V
Assume $h_{FE} = 50$ (typical value for HP transistors)

$$I_B = I_C/h_{FE} = 10 \text{ mA}/50 = .2 \text{ mA} \quad (3)$$

so,

$$R_B = \frac{V_{CE}-V_{BE}}{I_B} = \frac{10-.7}{.2 \text{ mA}} = 46.5 \text{ k}\Omega \quad (4)$$

and,

$$R_C = \frac{V_{CC}-V_{CE}}{I_C+I_B} = \frac{20-10}{10.2 \text{ mA}} = .98 \text{ k}\Omega \quad (5)$$

substituting standard resistor values,

$$R_B = 46.4 \text{ k}\Omega \quad (6)$$
$$R_C = 1 \text{ k }\Omega \quad (7)$$

Since $V_{CE} \sim 1/h_{FE}$, the base resistor, R_B, can be adjusted to compensate for manufacturing variations in h_{FE}. This is accomplished by varying R_B while monitoring V_{CE} to obtain the desired collector current. The DC bias circuit is shown in Figure 2.

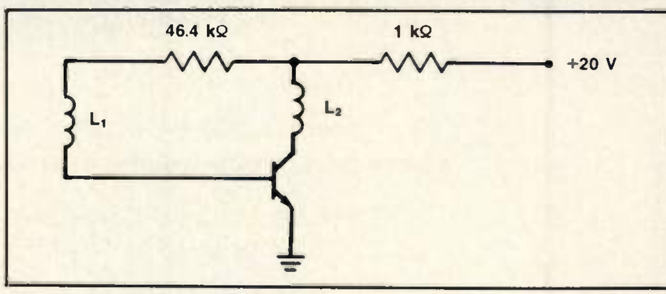

Figure 2. DC bias circuit.

L_1 and L_2 are RF chokes, and are used to isolate the DC bias network and the RF signal at the base and collector of the transistor. They should present a high reactance to the RF signal over the desired operating bandwidth relative to the source and load impedances.

$$X_{L1} \text{ and } X_{L2} >> 50 \ \Omega \quad (8)$$
$$X_{L1} = X_{L2} > (100)(50) = 5 \text{ k}\Omega \quad (9)$$

at 1 GHz,

$$X_{L2} = X_{L1} = \omega L_1 = \omega L_2 \quad (10)$$

$$L_1 = L_2 = \frac{X_{L1}}{\omega} = \frac{5 \times 10^3}{(2\pi)(1 \times 10^9)} = 795 \text{ nH} \quad (11)$$

The bias circuit with the calculated values are shown in the Figure 3.

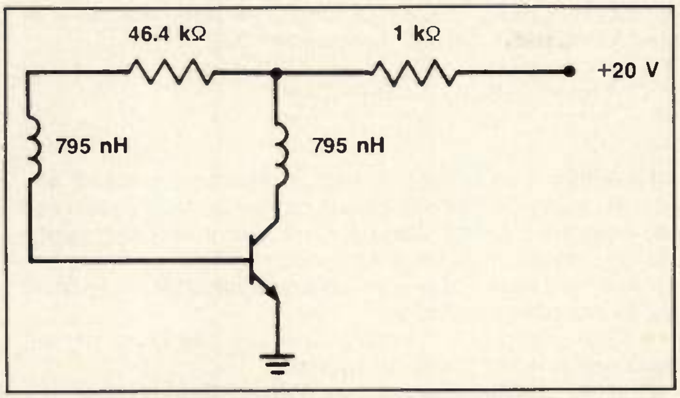

Figure 3. Completed transistor bias circuit.

Transistor S-Parameters

The common-emitter transistor circuit shown in Figure 3 is a two-port network and can be modeled conveniently using scattering or S-parameters. S-parameters are well suited for describing transistors and other active devices(5). Measuring other parameters, such as h, Z, and Y, requires the input and output of the device under test to be successively opened and short-circuited. This is difficult to do at microwave frequencies where lead inductance and capacitance make short and open circuit realization a somewhat challenging venture. S-parameters, on the other hand, are measured with the device under test imbedded between a 50Ω source and load, and the chance for oscillation is minimized(5).

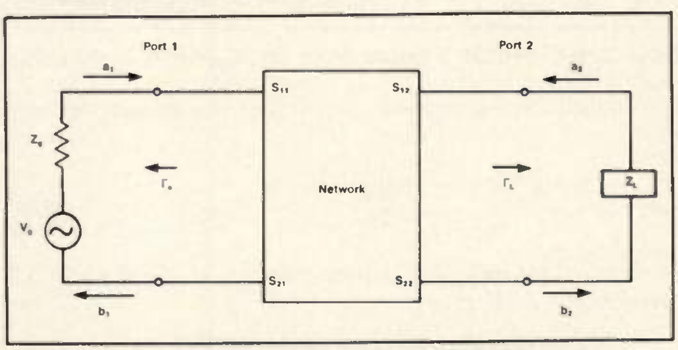

Figure 4. S-parameters for a two-port network.

Figure 4 shows the S-parameters for a two-port network. Define a_1 and a_2 as incident traveling waves on port 1 and port 2, respectively and b_1 and b_2 as traveling waves emanating from port 1 and port 2, respectively. These four quantities are related by equations 12 and 13.

$$b_1 = S_{11}a_1 + S_{12}a_2 \tag{12}$$

$$b_2 = S_{21}a_1 + S_{22}a_2 \tag{13}$$

Or, in matrix form:

$$\begin{bmatrix} b_1 \\ b_2 \end{bmatrix} = \begin{bmatrix} S_{11} & S_{12} \\ S_{21} & S_{22} \end{bmatrix} \begin{bmatrix} a_1 \\ a_2 \end{bmatrix} \tag{14}$$

so,

$S_{11} = \left.\dfrac{b_1}{a_1}\right| a_2 = 0$ ≐ Input reflection coefficient with port 2 terminated by a matched load.

$S_{22} = \left.\dfrac{b_2}{a_2}\right| a_1 = 0$ ≐ Output reflection coefficient with port 1 terminated by a matched load.

$S_{21} = \left.\dfrac{b_2}{a_1}\right| a_2 = 0$ ≐ Forward transmission gain with port 2 terminated in a matched load.

$S_{12} = \left.\dfrac{b_1}{a_2}\right| a_1 = 0$ ≐ Reverse transmission gain with port 1 terminated in a matched load.

also,

$|a_1|^2$ = Power incident on port 1 of network
= Power available from a source of impedance Z_o

$|a_2|^2$ = Power incident on port 2 of the network
= Power reflected from the load

$|b_1|^2$ = Power reflected from port 1

$|b_2|^2$ = Power emanating or reflected from port 2

$|S_{11}|^2 = \dfrac{\text{Power reflected from port 1}}{\text{Power incident on the network input}}$

$|S_{22}|^2 = \dfrac{\text{Power reflected from port 2}}{\text{Power incident on port 2}}$

$|S_{21}|^2 = \dfrac{\text{Power delivered to a } Z_o \text{ load}}{\text{Power available from a } Z_o \text{ source}}$
= Power gain with Z_o load and source

$|S_{12}|^2$ = Reverse power gain with Z_o load and source

The power gain (or loss) of the 2-port network, G_T, is defined as (5):

$$G_T = \frac{\text{Power delivered to the load}}{\text{Power available from the source}} \tag{15}$$

$$G_T = \frac{|S_{21}|^2(1-|\Gamma_s|^2)(1-|\Gamma_L|^2)}{|(1-S_{11}\Gamma_s)(1-S_{22}\Gamma_L)-S_{21}\,S_{12}\,\Gamma_L\,\Gamma_s\,|^2} \tag{16}$$

Equation 16 applies for general values of Γ_S and Γ_L. Γ_L is the reflection coefficient of the load relative to Z_o of the output transmission line. Γ_S is the reflection coefficient of the source relative to Z_0 of the input transmission line.

Many microwave transistors have "unilateral" properties. By this we mean that $|S_{12}| \approx 0$ or so small that it is negligible. S_{12} can be visualized as negative feedback internal to the transistor which allows the load reflection coefficient and some of its associated reflected power to propagate back to port 1, thus influencing the gain of the overall network. If $|S_{12}| \approx 0$, then the power reflected from the load is in effect isolated from port 1 (base of transistor). When this is the case, it reduces equation 16 to the simpler equation 17 for unilaterial power gain (G_{TU})(4).

$$G_{TU} = \left[\frac{(1-|\Gamma_s|^2)}{|1-S_{11}\Gamma_s|^2}\right]\left[|S_{21}|^2\right]\left[\frac{(1-|\Gamma_L|^2)}{|1-S_{22}\Gamma_L|^2}\right] = (G_s)(G_o)(G_L) \tag{17}$$

Equation 17 is made up of three distinct and independent gain factors and the amplifier as three distinct gain blocks as shown in Figure 5 (4).

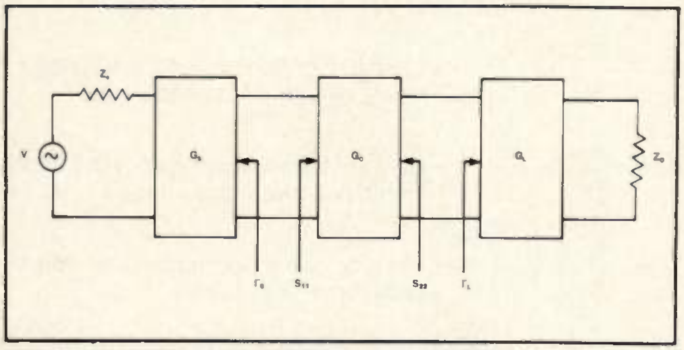

Figure 5. Unilateral transistor amplifier model.

The factor G_o of equation 17 is related to the transistor only. Once the device and its bias conditions are established, S_{21} is determined and remains invariant throughout the design.

The factor G_S of equation 17 affects the degree of mismatch between the characteristic impedance of the source and the input reflection coefficient of the transistor. Even though the block G_S is made up of passive components, it can have a gain contribution greater than unity. This is true because an intrinsic mismatch exists between Z_o and S_{11} and impedance transforming elements can be employed to improve this match, thus decreasing the mismatch loss, and can be thought of as providing gain (4).

The third term of equation 17, G_L, serves the same function as the G_S term, but affects the matching at the output rather than the input.

Maximum unilateral gain can be accomplished by choosing impedance matching networks such that $\Gamma_S = S_{11}{}^*$ and $\Gamma_L = S_{22}{}^*$. Equation 17 then reduces to equation 18 and is graphically depicted in Figure 6 (4).

$$G_{TUmax} = \left[\frac{1}{1-|S_{11}|^2}\right]\left[|S_{21}|^2\right]\left[\frac{1}{1-|S_{22}|^2}\right] \tag{18}$$

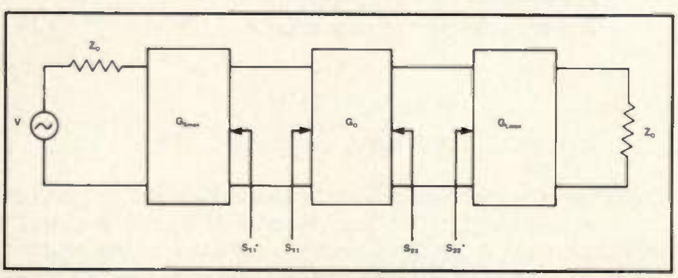

Figure 6. Maximized gain model for unilateral amplifier.

Transistor Stability

Another consideration in the amplifier design process is stability. A network is conditionally stable if the real part of Z_{in} and Z_{out} is greater than zero for some positive real source and load impedance at a specific frequency (4).

A network is unconditionally stable if the real part of Z_{in} and Z_{out} is greater than zero for all positive real source and load impedances at a specified frequency (4).

Unconditional stability is a desired attribute of a transistor, especially for broadband design. This is because Γ_s and Γ_L are not constant with frequency, but vary linearly over the operating frequency range (for ideal components).

Fortunately, unconditional stability can be checked by calculating the Rollett stability factor K(2). If K is greater than 1 at a specific frequency, and the transistor is biased properly, the transistor will be stable for all passive values of Γ_S and Γ_L. The Rollett stability factor is a function of the transistor's S-parameters and is defined in equation 19(2).

$$K = \frac{1+|S_{11}S_{22}-S_{12}S_{21}|^2-|S_{11}|^2-|S_{22}|^2}{2|S_{12}|\,|S_{21}|} > 1 \tag{19}$$

If K is less than 1, the transistor is potentially unstable and will most likely oscillate with certain combinations of source and load impedance. Extreme care must be taken in choosing source and load impedances for the transistor.

If K is less than 1, there are several approaches one could take to complete the design:

(1) forge ahead with the design and see how lucky you are (good way to end up with an oscillator).

(2) select another bias point for different S-parameters.

(3) choose a different transistor.

(4) calculate acceptable values for Γ_S and Γ_L (difficult for broadband design).

The problem of potential instability has been avoided in this design by using HP HXTR-3101 transistor in the 1-2 GHz frequency band.

At this point we are now ready to calculate the Rollett stability factor for the transistor. The S-parameters are found on the transistor data sheet and shown in Table 1.

1.0 GHz	1.5 GHz	2.0 GHz
S_{11}=.571$\angle$-169°	S_{11}=.574$\angle$174°	S_{11}=.591$\angle$161°
S_{12}=.052$\angle$44°	S_{12}=.066$\angle$48°	S_{12}=.080$\angle$48°
S_{21}=5.33$\angle$78°	S_{21}=3.627$\angle$63°	S_{21}=2.774$\angle$49°
S_{22}=.408$\angle$-43°	S_{22}=.394$\angle$-48°	S_{22}=.392$\angle$-57°

Table 1. HXTR-3101 S-parameters for V_{CE}=10 V, I_C=10 mA.

Substituting the S-parameters for 1.0 GHz into equation 19, we obtain:

$$K = \frac{1+(.3620659)^2 - (.571)^2 - (.408)^2}{(2)(.052)(5.33)} = 1.152 \tag{20}$$

Since K > 1, the transistor is unconditionally stable at 1.0 GHz. Likewise, at 1.5 GHz:

$$K = \frac{1+(.0621637)^2 - (.574)^2 - (.394)^2}{(2)(.066)(3.627)} = 1.084 \tag{21}$$

Since K > 1, the transistor is unconditionally stable at 1.5 GHz. Finally, at 2.0 GHz:

$$K = \frac{1+(.0293521)^2 - (.591)^2 - (.392)^2}{(2)(.08)(2.774)} = 1.121 \tag{22}$$

Again K > 1, the transistor is unconditionally stable at 2.0 GHz.

We have now established unconditional stability for the HXTR-3101 transistor at 1.0, 1.5, and 2.0 GHz, provided the transistor is biased at V_{CE} = 10 V, and I_C = 10 mA. The best way to prove the transistor is unconditionally stable through the 1-2 GHz band is to test the transistor with a microwave vector network analyzer and a digital computer. For the purposes of this design, assume unconditional stability based on the transistor's calculated stability at 1.0, 1.5, and 2.0 GHz.

The source and load matching networks will provide an overall amplifier circuit power gain of 10 dB while at the same time maintaining a noise figure of less than 3.5 dB.

Noise Figure Circles

The noise figure, or "F" of a network, is a quantity used as a figure of merit to compare the noise in a network with the noise in an ideal or lossless network (2). It is a measure of the degradation in signal-to-noise ratio (SNR) between the input and output ports of the network.

$$F = 10 \operatorname{Log}_{10} \frac{\text{Input SNR}}{\text{Output SNR}} \text{ dB} \quad (23)$$

Using noise parameters, the noise figure of a single-stage transistor amplifier is a function of Γ_s — the reflection coefficient of the input matching network (6).

$$F(\Gamma_s) = F_{min} + 4r_n \frac{|\Gamma_s - \Gamma_o|^2}{(1 - |\Gamma_s|^2)|1+\Gamma_o|^2} \quad (24)$$

where,

F_{min} = Minimum Noise Figure

$r_n = \dfrac{R_n}{50}$ = Normalized equivalent noise resistance

Γ_o = Optimum source reflection coefficient for F_{min}

Notice that if $\Gamma_s = \Gamma_o$, the second term of equation 24 goes to zero, and equation 24 reduces to $F = F_{min}$. F_{min}, r_n, and Γ_o are noise parameters of the transistors. They are usually supplied by the manufacturer and vary with frequency and DC bias point.

The noise parameters obtained from HP for the HXTR-3101 are shown in Table 2 (3).

Freq. (GHz)	F_{min}	Γ_o	R_n
.5	1.4 dB	$.121\angle 96°$	114.4Ω
1.0	1.7 dB	$.301\angle 121°$	15.2Ω
2.0	2.5 dB	$.461\angle 173°$	5.2Ω
3.0	3.3 dB	$.553\angle -157°$	8.4Ω
4.0	4.2 dB	$.648\angle -139°$	13.4Ω

Table 2. Noise Parameters for HXTR-3101 (V_{CE} = 10 V, I_C = 10 mA)

Recall that the design goals for this amplifier design are:

Bandwidth: 1-2 GHz
Noise Figure: less than 3.5 dB
Gain: 10 dB (small-signal)

Clearly, Γ_s for the input matching network used in conjunction with Γ_L for the output matching network to achieve 10 dB of gain will be constrained by equation 24, such that:

$$F < 3.5 \text{ dB} \quad (25)$$

In order to determine acceptable values of Γ_s that will meet the constraint of equation 25, it is necessary to establish the 3.5 dB noise figure circles a 1 GHz and 2 GHz. These noise figure circles can be plotted on a smith chart to simplify the design process. Once plotted, all values of Γ_s within the boundaries of the noise circles will provide an overall amplifier noise figure of less than 3.5 dB.

To determine the 1 GHz and 2 GHz 3.5 dB noise figure circles, equations 26, 27, and 28 can be used (6).

$$N_i = \left[\frac{F_i - F_{min}}{4r_n}\right] |1+\Gamma_o|^2 \quad (26)$$

$$C_i = \frac{\Gamma_o}{1 + N_i} \quad (27)$$

$$R_i = \frac{\sqrt{N_i^2 + N_i(1-|\Gamma_o|^2)}}{1 + N_i} \quad (28)$$

F_i is the value of the desired noise figure circle. Γ_o, F_{min}, and r_n are the noise parameters of the transistor. C_i is the location on the Smith chart of the center of the noise figure circle and R_i is the radius of noise figure circle. N_i is an intermediate quantity used in equations 27 and 28. Substituting the noise parameters in Table 2 at 1 GHz into equation 26, at 1 GHz:

$$N_i = \left[\frac{3.5 - 1.7}{(4)\frac{15.2}{50}}\right] |1 + .301\angle 121°|^2 = 1.1554166 \quad (29)$$

Likewise, substituting N_i and Γ_o for 1 GHz into equation 27 at 1 GHz:

$$C_i = \frac{.301\angle 121°}{2.1554166} = .1396\angle 121° \quad (30)$$

C_i, the center point, is plotted on the Smith chart in the same manner as a reflection coefficient. The radius of the Smith chart used in this design is 11.4 cm. Thus C_i multiplied by 11.4 is the quantity plotted on the Smith chart at 1 GHz.

$$(11.4)C_i = 1.591\angle 121° \text{ cm} \quad (31)$$

To find the radius of the 3.5 dB noise figure at 1 GHz, substitute the value of N_i in equation 29 and Γ_o from Table 2 into equation 28.

At 1 GHz,

$$R_i = \frac{\sqrt{(1.1554166)^2 + (1.1554166)(1 - (.301)^2)}}{2.1554166} \quad (32)$$

$R_i = .7166$

R_i, calculated in equation 32, is a fractional number between 0 and 1 which represents the radius of the noise figure circle relative to the radius of the Smith chart. Since we are using a Smith chart with a radius of 11.4 cm, at 1 GHz,

$$\text{Radius of circle} = (11.4)R_i = 8.169 \text{ cm} \quad (33)$$

Following the same procedure at 2 GHz,

$$N_i = \left[\frac{3.5 - 2.5}{(4)\frac{5.2}{50}}\right] |1 + .461\angle 173°|^2 \quad (34)$$

$N_i = .7148881$

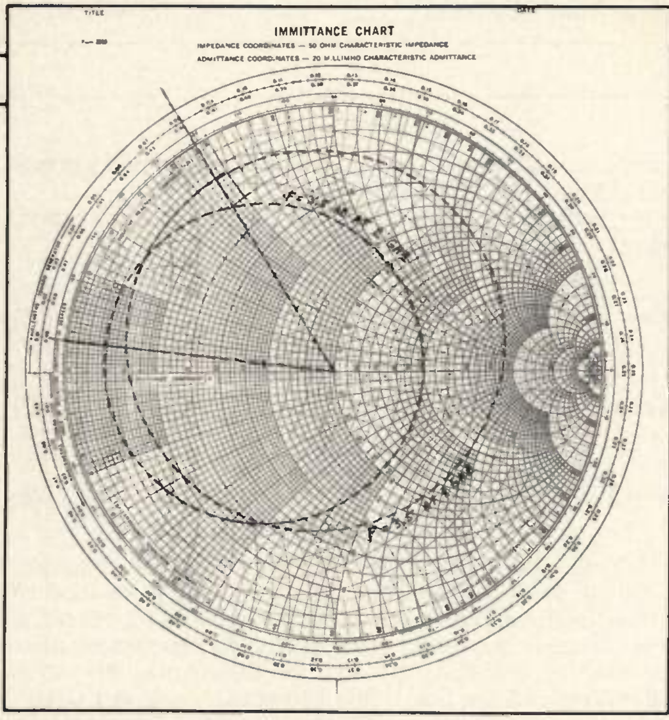

Figure 7. Noise figure circles.

$$C_i = \frac{.461\angle 173^\circ}{1.7148881} = .2688\angle 173^\circ \quad (35)$$

$$(11.4)(C_i) = 3.96\angle 173^\circ \text{ cm} \quad (36)$$

$$R_i = \frac{\sqrt{(.7148881)^2 + (.7148881)[1 - (.461)^2]}}{1.7148881} \quad (37)$$

$$R_i = .6043$$

$$(11.4)R_i = 6.88 \text{ cm} \quad (38)$$

A plot of the 3.5 dB noise figure circles at 1 GHz and 2 GHz is shown in Figure 7. As stated before, the input matching network reflection coefficient, Γ_s, must be located within these circles at 1.0 and 2.0 GHz to retain an overall amplifier noise figure of less than 3.5 dB.

Constant Gain Circles

With the aid of a Smith chart, note that there are 3.5 dB noise figure circles at 1 and 2 GHz. The task remaining is to calculate values for the passive elements used in the input and output matching networks that will provide 10 dB of power gain for the amplifier from 1 to 2 GHz.

To facilitate this task, assume that the HXTR-3101 transistor is unilateral, that is, the value of $|S_{12}|$ is small enough to be neglected. In doing so, realize that a small error will be introduced in the overall gain of the amplifier. This error is small enough to be ignored since the S-parameters on the datasheet are strictly typical values anyway.

Based on the unilateral transistor assumption above, the following steps will be used for designing the input and output matching networks:

(1) calculate the maximum unilateral gain of the amplifier at 1 and 2 GHz (band edges) using equation 18 and the values in Table 1.
(2) calculate the required gain of the input and output matching networks to achieve an overall amplifier gain of 10 dB at 1 and 2 GHz.
(3) calculate and plot on a Smith chart the input and output constant gain circles at 1 and 2 GHz (using gain values found in step 2).
(4) determine values of the elements in the input and output matching networks using the Smith chart and broadband impedance matching techniques.
(5) calculate the amplifier gain at 1.5 GHz (center frequency).

Equation 18 can be rewritten as:

$$G_{TUmax}(dB) = 10 \text{ Log}\left[\frac{1}{1-|S_{11}|^2}\right] + 10 \text{ Log } |S_{21}|^2$$

$$+ 10 \text{ Log}\left[\frac{1}{1-|S_{22}|^2}\right] = G_{Smax}(dB) + G_o(dB) + G_{Lmax}(dB) \quad (39)$$

Substituting the values in Table 1 for 1 GHz into equation 39:

$$G_{TUmax}\text{ (1 GHz)} = 1.7136 \text{ dB} + 14.5345 \text{ dB} + .79075 \text{ dB} = 17.0389 \text{ dB} \quad (40)$$

The first term in equation 40 indicates the maximum gain contribution to the amplifier due to input matching is 1.7136 dB. The second term is the transistor gain which is independent of input and output matching and is equal to 14.5345 dB. The third term of equation 40 is the maximum gain contribution to the amplifier due to output matching and is .79075 dB. The transistor gain at 1 GHz with no matching at all is over 14 dB. This means that at 1 GHz, the input and output matching networks need to provide negative gain (or attenuation) for the overall amplifier gain to be 10 dB.

Calculating required values for G_S and G_L at 1 GHz, G_o = 14.5345 dB

Desired $G_{TU} = 10 \text{ dB} = G_0 + G_S + G_L$

so,

$$G_S + G_L = G_{TU} - G_o = 10 \text{ dB} - 14.5345 \text{ dB}$$

$$G_S + G_L = -4.5345 \text{ dB} \quad (41)$$

Equation 41 indicates that at 1 GHz, the combined gain contribution of the input and output matching networks is equal to −4.5345 dB. For simplicity set $G_S = G_L$ and divide the −4.5345 dB of gain equally between the two. Then, at 1 GHz,

$$G_S = -2.2672 \text{ dB} \quad (42)$$
$$G_L = -2.2672 \text{ dB} \quad (43)$$

Following the same procedure at 2 GHz:

$$G_{TUmax} = G_{Smax}(dB) + G_0(dB) + G_{Lmax}(dB) = 1.866 \text{ dB} + 8.8621 \text{ dB} + .7245 \text{ dB} = 11.4527 \text{ dB} \quad (44)$$

At 2 GHz, the transistor gain, G_o, is only 8.8621 dB. This means that at 2 GHz, the input and output matching networks need to provide positive gain for the overall amplifier gain to be 10 dB.

Calculating the required values for G_S and G_L at 2 GHz,

G_o = 8.8621 dB

Desired $G_{TU} = 10 \text{ dB} = G_o + G_S + G_L$

so,

$G_S + G_L = 10\text{ dB} - G_o = 10\text{ dB} - 8.8621\text{ dB}$
$G_S + G_L = 1.1378\text{ dB}$ (45)

Again, as was done at 1 GHz, for simplicity set $G_S = G_L$ and divide the 1.1378 dB of gain equally between the two.

Then, at 2 GHz,
$G_S = .5689\text{ dB}$ (46)
$G_L = .5689\text{ dB}$ (47)

Table 3 shows the calculated input and output matching network gain figures in dB at 1.0 and 2.0 GHz.

	G_o(dB)	G_S(dB)	G_L(dB)	G_{TU}(dB)
1 GHz	14.5345	−2.2672	−2.2672	10.0 dB
2 GHz	8.8621	0.5689	0.5689	10.0 dB

Table 3. Matching gain figures.

Equations 48, 49 and 50 are used to plot the constant gain circles for the input matching network (4).

$$g_s = \frac{G_S}{G_{Smax}} \quad (G_S \text{ not in dB}) \qquad (48)$$

$$d_s = \frac{g_s|S_{11}|}{1 - |S_{11}|^2(1 - g_s)} \qquad (49)$$

$$R_s = \frac{\sqrt{1 - g_s}\,(1 - |S_{11}|^2)}{1 - |S_{11}|^2(1 - g_s)} \qquad (50)$$

The quantity d_s is the distance from the center of the Smith chart to the center of the constant gain circle along the vector $S_{11}{}^*$ and is plotted in the same manner as a reflection coefficient. R_s is a fractional number which corresponds to the radius of the constant gain circle relative to the radius of the Smith chart. The quantity g_s is the normalized gain value for the G_S constant gain circle (not in dB).

Substituting values from Table 1 and Table 3 into equations 48, 49 and 50, at 1 GHz:

$G_S = -2.2672\text{ dB} = .59329$

$G_{Smax} = 1.7136\text{ dB} = 1.48377$

$$g_s = \frac{.59329}{1.48377} = .399858 \qquad (51)$$

$d_s = .28386$ (52)

$(11.4\text{ cm})\, d_s = 3.236\text{ cm (along } 169°)$ (53)

$R_S = .649122$ (54)

$(11.4\text{ cm})\, R_s = 7.399\text{ cm}$ (55)

at 2 GHz,

$G_S = .5689\text{ dB} = 1.139935$

$G_{Smax} = 1.8660\text{ dB} = 1.53676$

$$g_s = \frac{1.139935}{1.53676} = .7418003 \qquad (56)$$

$d_s = .4818601$

$(11.4\text{ cm})d_s = 5.493\text{ cm (along } -161°)$ (57)

$R_s = .363427$

$(11.4\text{ cm})R_s = 4.143\text{ cm}$ (58)

Figure 8 shows the input matching constant gain circles at 1.0 and 2.0 GHz. For convenience, the constant noise figure circles of Figure 7 are superimposed. Any impedance network which transforms the 50 Ω source impedance to a point *on* the constant gain circles at 1 and 2 GHz, respectively, and also stays inside the 3.5 dB noise figure circles is acceptable for the input matching circuit. Shown on Figure 8 are the input impedance transform arcs. An ideal shunt inductor (arc A-C) has a constant conductance of zero and its susceptance decreases with frequency. The arc-length of A-C is equal to the susceptance of the shunt inductor at 1 GHz. The length of arc A-B on the Smith chart is equal to the susceptance of the shunt inductor at 2 GHz.

An ideal series capacitor (arc C-D and B-E) has a constant resistance of zero and its reactance decreases with frequency. The arc-length of C-D on the Smith chart is equal to the reactance of the series capacitor at 1 GHz. The length of arc B-E is the reactance of the series capacitor at 2 GHz.

It is obvious that with the right values of shunt inductance and series capacitance, the input impedance transforming (matching) network can *land* on the required constant gain circles. Using the impedance transform arcs in Figure 8, the values of the shunt inductor and the series capacitor can now be calculated:

	1 GHz	2 GHz
shunt L,	arc A-C=20 millimhos	arc A-B=10 millimhos
series C,	arc C-D=64 ohms	arc B-E=32 ohms

Now calculating the value of the shunt inductor and series capacitor
at 1 GHz, $B_L = .02$ mhos
at 2 GHz, $B_L = .01$ mhos

$$L = \frac{1}{\omega B_L} = \frac{1}{(2\pi)(1\times10^9)(.02)} = \frac{1}{(2\pi)(2\times10^9)(.01)} \qquad (59)$$

$= 7.95\text{ nH}$

at 1 GHz, $X_c = 64\Omega$
at 2 GHz, $X_c = 32\Omega$

$$X_c = \frac{1}{\Omega C}$$

$$C = \frac{1}{\omega X_c} = \frac{1}{(2\pi)(1\times10^9)(64)} = \frac{1}{(2\pi)(2\times10^9)(32)} = 2.48\text{pF} \qquad (60)$$

We now know the value of the shunt inductor and the series capacitor which form the amplifier's input matching circuit.

All that remains now is to find the required output matching circuit to produce the values of G_L in Table 3 at 1 and 2 GHz and the amplifier design is completed. Equations 61, 62 and 63 are used to plot the constant gain circles for the output matching network (4).

$$g_L = \frac{G_L}{G_{Lmax}} \quad (G_L \text{ not in dB}) \qquad (61)$$

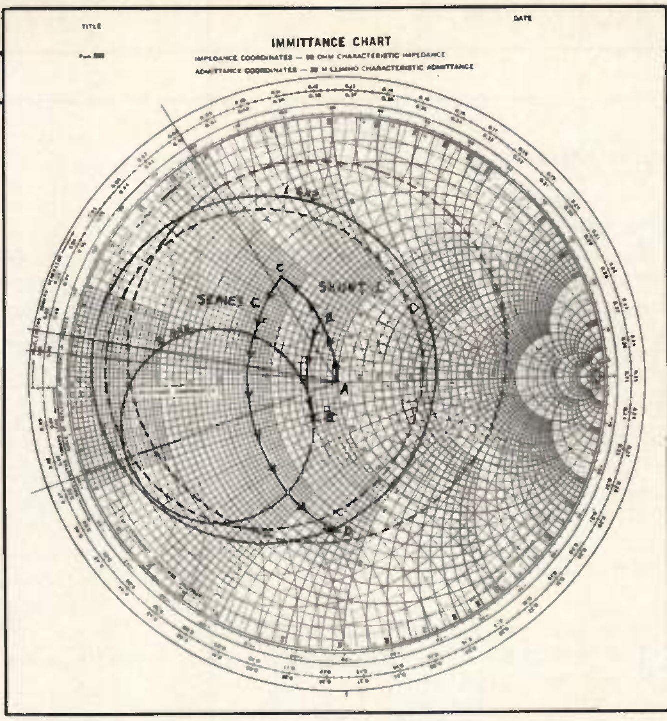

Figure 8. Noise figure and input matching.

$$d_L = \frac{g_L|S_{22}|}{1 - |S_{22}|^2(1 - g_L)} \tag{62}$$

$$R_L = \frac{\sqrt{1 - g_L}\,(1-|S_{22}|^2)}{1 - |S_{22}|^2(1 - g_L)} \tag{63}$$

The quantities g_L, d_L, and R_L have the same meaning as g_s, d_s, and R_s except that they are referenced to the output matching network.

Substituting values from Table 1 and Table 3 into equations 61, 62 and 63, we obtain:

at 1 GHz,

G_L = −2.2672 dB = .59329
G_{Lmax} = .79075 dB = 1.1997

$$g_L = \frac{G_L}{G_{Lmax}} = .494535 \tag{64}$$

d_L = .220307 (65)
(11.4 cm)d_L = 2.511 cm (along 43°) (66)
R_L = .647055 (67)
(11.4 cm)R_L = 7.376 cm (68)

at 2 GHz,

G_L = .5689 dB = 1.13997
G_{Lmax} = .7245 dB = 1.1815

$$g_L = \frac{G_L}{G_{Lmax}} = .964797 \tag{69}$$

d_L = .380257 (70)
(11.4 cm)d_L = 4.334 cm (along 57°) (71)
R_L = .159655 (72)
(11.4 cm)R_L = 1.820 cm (73)

The output matching circuit constant gain circles are plotted on a Smith chart in Figure 9 along with the impedance transfor-

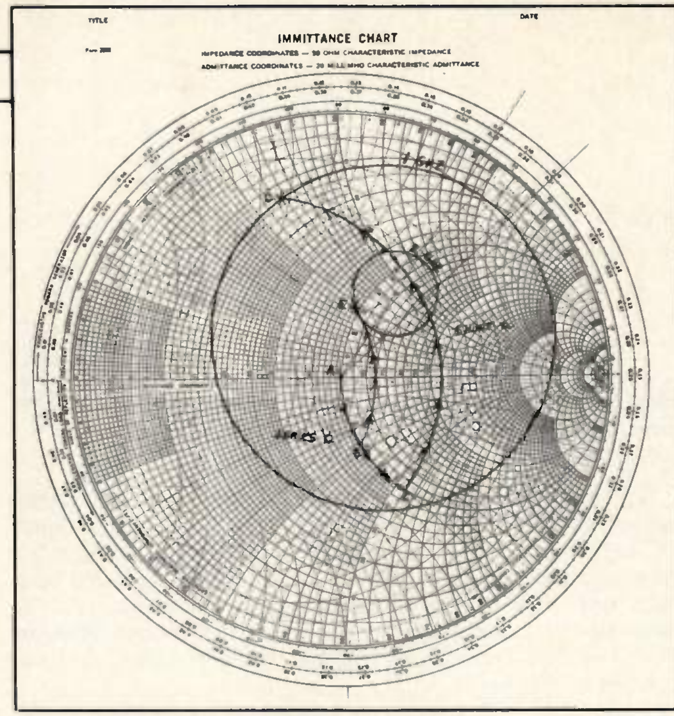

Figure 9. Output matching circles.

ming arcs. As in the input matching plot, we again need a series capacitor and a shunt indicator. By examining the arcs on the Smith chart, we obtain:

	1 GHz	2 GHz
Series C,	arc A-C = 54Ω	arc A-B = 27Ω
Shunt L,	arc C-D=35 millimhos	arc B-E=17.5 millimhos

We can now calculate the value of the series capacitor and the shunt inductor.

at 1 GHz, X_c = 54Ω
at 2 GHz, X_c = 27Ω

$$X_c = \frac{1}{\omega C}$$

$$C = \frac{1}{\omega X_c} = \frac{1}{(2\pi)(1\times10^9)(54)} = \frac{1}{(2\pi)(2\times10^9)(27)} = 2.94 \text{ pF} \tag{74}$$

at 1 GHz, B_L = .035 mhos
at 2 GHz, B = .0175 mhos

$$B_L = \frac{1}{\omega L}$$

$$L = \frac{1}{\omega B_L} = \frac{1}{(2\pi)(1\times10^9)(.035)} = \frac{1}{(2\pi)(2\times10^9)(.0175)} = 4.54 \text{ nH} \tag{75}$$

Table 4 shows the values of the input and output matching components and Figure 10 shows a complete schematic of the amplifier.

	Shunt L	Series C
input circuit	7.95 nH	2.48 pF
output circuit	4.54 nH	2.94 pF

Table 4. Amplifier input and output matching components.

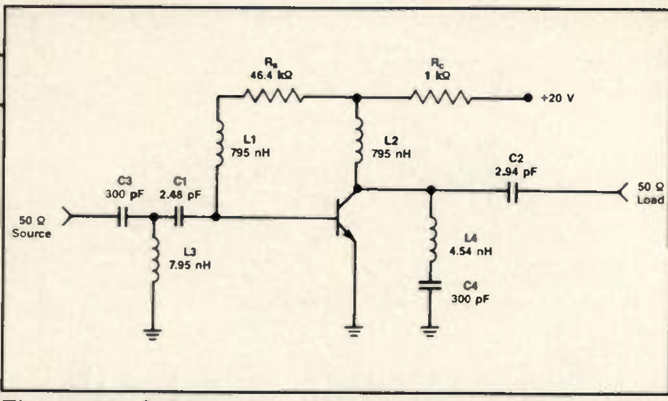

Figure 10. Schematic of completed amplifier design with bias and matching circuitry.

Notice the addition of C_3 and C_4 to the circuit. These capacitors are not part of the matching circuitry. C_3 is an input DC blocking capacitor implemented to protect L_3 in case a DC voltage is applied to the input of the amplifier. C_3 should be a sufficiently high value of capacitance such that its reactance is much smaller than that of the input matching components within the operating bandwidth, so its effect on the input matching circuit will be negligible.
at 1 GHz,

$$X_{C3} << X_{C1} = \frac{1}{\omega C_1} = 64.17\Omega$$

$$X_{C3} << X_{L3} = \omega L_3 = 49.95\Omega$$

If we choose C_3 = 300 pF, then

$$X_{C3} = .53\Omega = \frac{X_{C1}}{121} = \frac{X_{L3}}{94}$$

This justifies 300 pF for the value of C_3.

C_4 is needed to keep the collector DC bias current from being shorted to ground through L_4, the output matching shunt inductor. Since C_4 is in series with L_4, its reactance should be much less than that of L_4 over the operating bandwidth, so that C_4 will have a negligible effect on the output matching circuit.
at 1 GHz,

$$X_{C4} << X_{L4} = \omega L_4 = 28.52\Omega$$

If we choose C_4 = 300 pF, then:

$$X_{C4} = .53\Omega = \frac{X_{L4}}{53.8}$$

This justifies 300 pF for the value of C_4.

The gain at the center frequency (1.5 GHz) should now be checked using equation 17.

$$G_{TU} = \left[\frac{(1 - |\Gamma_s|^2)}{|1 - S_{11}\Gamma_s|^2}\right]\left[|S_{21}|^2\right]\left[\frac{(1 - |\Gamma_L|^2)}{|1 - S_{22}\Gamma_L|^2}\right]$$

Using interpolation on the Smith charts of Figures 8 and 9 to determine the values of Γ_s and Γ_L at 1.5 Ghz:

$\Gamma_s = .3508\angle{-112^\circ}$
$\Gamma_L = .5087\angle{91.5^\circ}$

Substituting the S-parameters for 1.5 GHz from Table 1, Γ_s, and Γ_L into equation 17 we obtain:

G_{TU}(1.5 GHz) = 13.339 = 11.271 dB

Therefore, the amplifier gain is slightly higher at the center frequency than at the 1.0 and 2.0 GHz band edges.

About the Author

Jackie Hughes obtained his BSEE from the University of Alabama in 1980. He is a senior engineer with Spectradyne, 1501 North Plano Road, P.O. Box 830775, Richardson, TX 75083-0775, Tel: (214) 234-2721 ext. 173.

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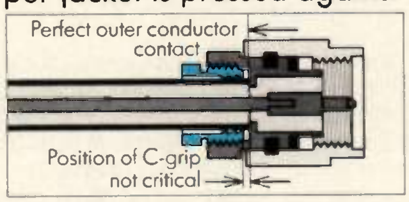

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VHF Oscillators Using Microwave Integrated Circuits

By Wes Hayward
TriQuint Semiconductor

Once considered a lab curiosity, the microwave integrated circuit (MIC) has become a common part of many RF systems. The most common MIC is the cascadable amplifier. These use negative feedback in a circuit containing one to perhaps a half dozen transistors to form an active two-port network with flat gain and well defined terminal impedances. The first MICs available were hybrid circuits, such as the Motorola MWA-120 series using a single transistor in a feedback amplifier. The monolithic microwave integrated circuit (MMIC) is now more common, with devices available from numerous vendors. The most popular topology is a Darlington pair, operating in a common-emitter configuration with negative feedback (1). This article shows how to use MICs in oscillators.

The MIC cascadable amplifier is typically used as a small signal or buffer circuit. It may also be used as the gain element in an oscillator. Although simple, the MIC amplifiers differ from the usual transistors used in an oscillator. The MIC cascadable amplifier usually contains negative feedback presenting moderate impedances (often 50 ohms) at the input and output ports.

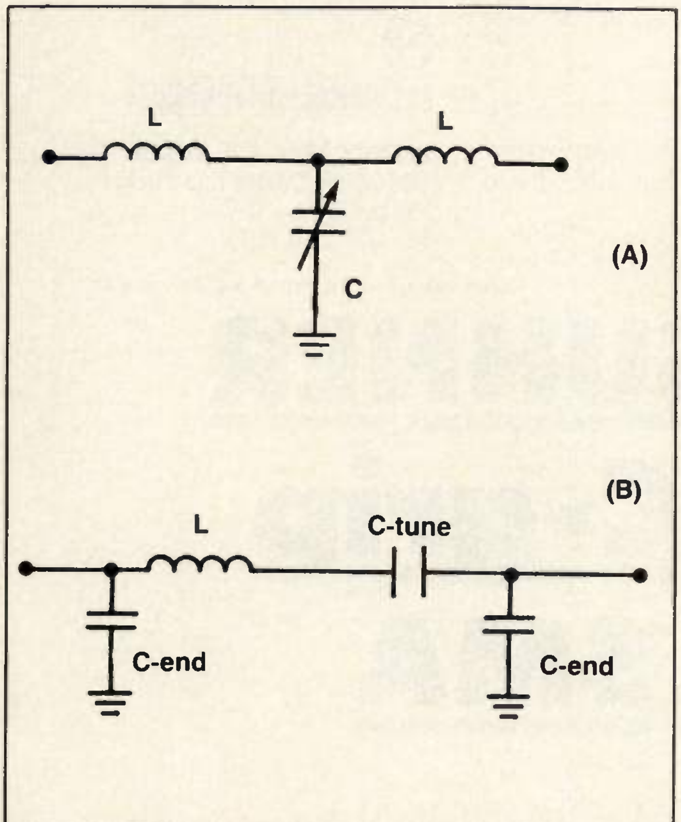


Figure 1. Resonant network with phase shift of 180°.

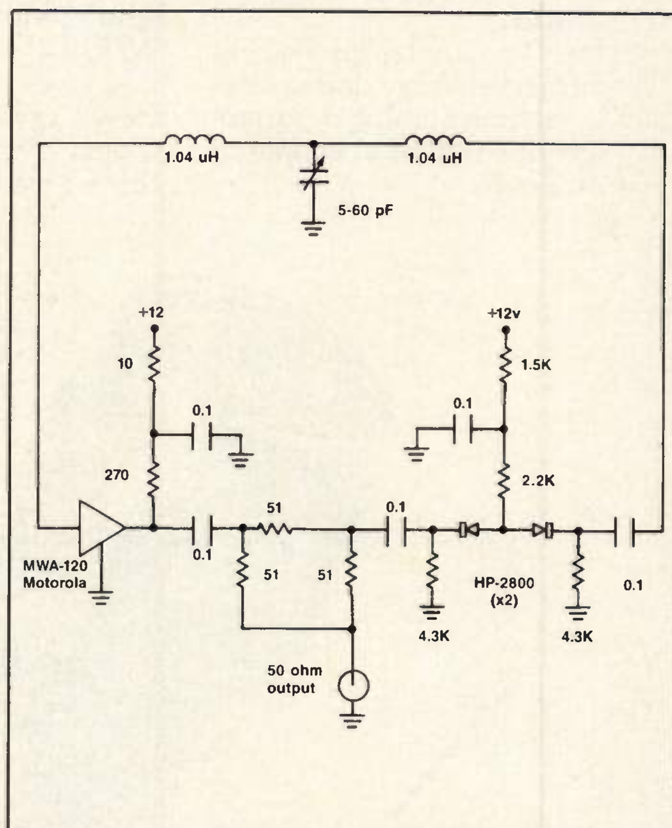


Figure 2. Experimental 50 MHz MIC oscillator.

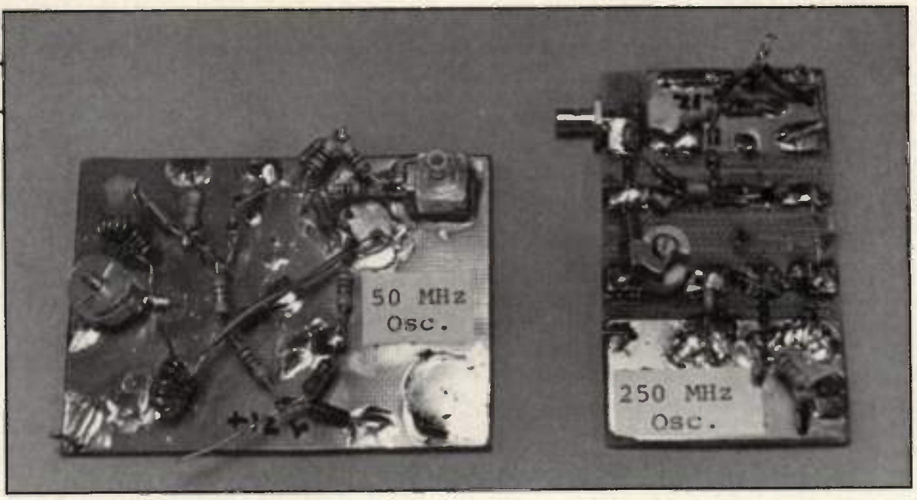


Photo of MMIC oscillators.

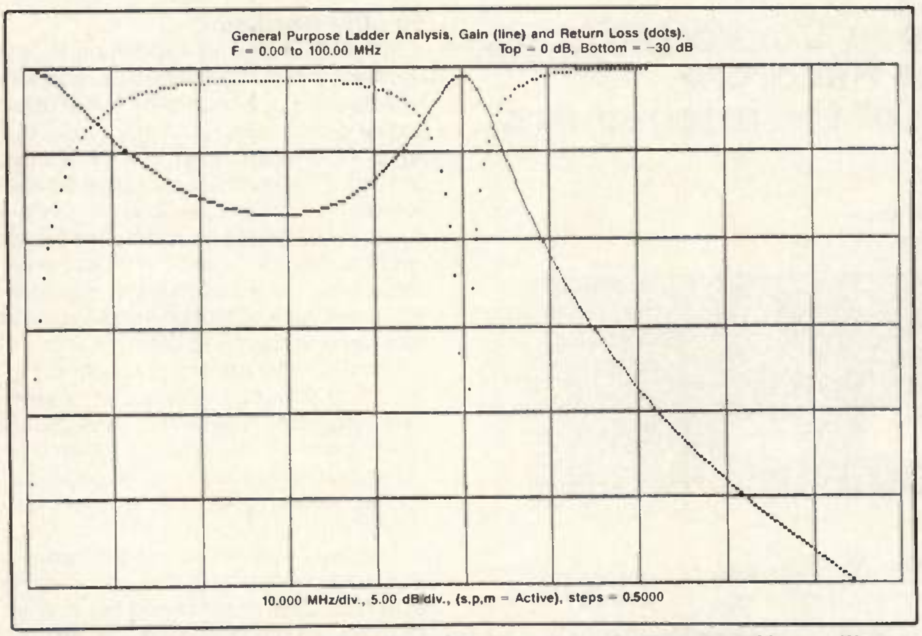


Figure 3. Resonator amplitude versus frequency for the 50 MHz oscillator.

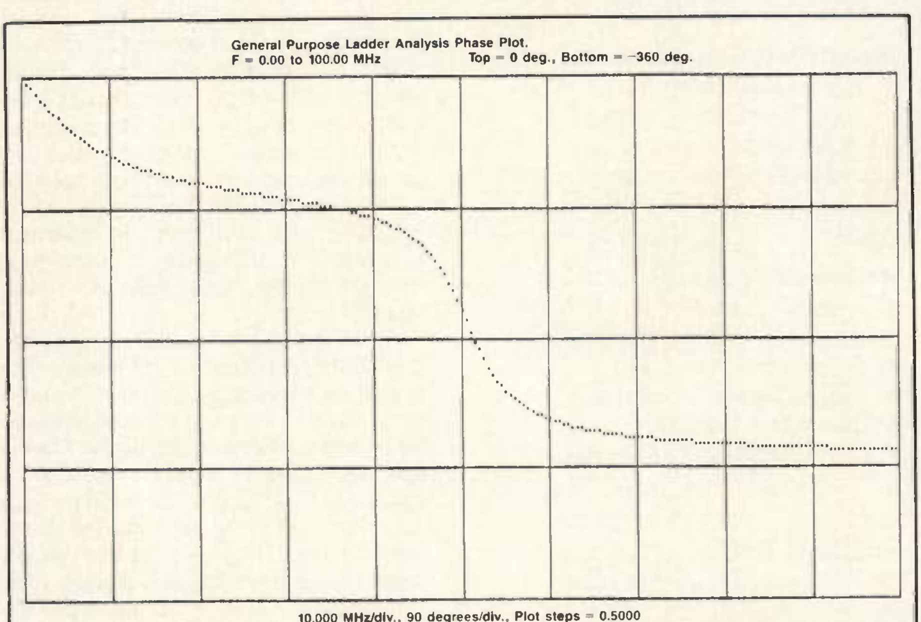


Figure 4. Plot of phase response for the 50 MHz oscillator.

An oscillator must contain a mechanism to limit its amplitude. This usually results from current limiting with the traditional bipolar transistor oscillator (2). The negative feedback built into the cascadable MIC amplifier usually precludes simple current limiting. MIC oscillator limiting external to the amplifier will ensure predictable performance.

Presently available MICs operate as inverting amplifiers. The angle of S21 is close to 180 degrees over much (usually, the lower half) of the amplifier bandwidth. A suitable oscillator will result when the amplifier is cascaded with a resonant network with a phase shift of 180 degrees at the frequency of peak amplitude response. Figure 1 shows two networks that can be built to generate such a phase response. Other networks, both lumped and distributed, have also been investigated and found suitable. Detailed analysis will depend upon available S-parameter data. If the circuit topology is known, it is possible to estimate S-parameters using a simple hybrid-pi transistor model. While less than ideal, a starting point is still provided. Analysis of the total circuit, even if approximate, is then possible.

Oscillator design may be performed by viewing the circuit as either a reflection amplifier, analyzed with S-parameters (3) or a two-port network that is analyzed with any convenient parameter set. The two-port approach is more intuitive when using the MIC cascadable amplifiers.

50 MHz Oscillator

Figure 2 shows an experimental MIC oscillator. A Motorola hybrid was used with limiting done by a back-to-back pair of hot carrier diodes biased to 1 mA. The circuit limits the signal current to a peak value of 1 mA. This is confirmed with a 50 ohm terminated oscilloscope. Assume that the fundamental driving frequency has a peak amplitude equal to the peak of the limited value, which is a square wave. The maximum available output power at the input frequency is then −16 dBm.

The resonator chosen is shown in Figure 1A. It was built with a pair of small toroid inductors, each with a value of 1.04 μH (17 turns on Micrometals T30-6). The unloaded Q was over 100 at 50 MHz, producing a network with an insertion loss of less than 1 dB.

The diode limiter produces an oscillator with predictable output power. Assuming a loss of 1 dB in the resonator, and an amplifier insertion gain of 13 dB, the available power at the amplifier output is −4 dBm. If all impedances were matched, the output power at the coax connector would be −10 dBm. The three 51-ohm resistors form a 6 dB power divider. However, if the input to the limiter becomes a high impedance during operation, the power available at the connector could be as high as −6.5 dBm. It was measured at −7 dBm with a homemade spectrum analyzer. The indicated accuracy is good considering the status of the instrumentation.

Figure 3 shows the calculated resonator amplitude response vs. frequency. The phase response is presented in Figure 4, while Figure 5 shows input impedance vs. frequency in a Smith chart format. The circuit was tuned for a peak amplitude response at 50 MHz (4). The oscillator was tuned from 24 to 90 MHz and the phase began to depart significantly from the desired 180 degrees at lower frequencies. Calculations indicate that the upper frequency is limited by parallel capacitance across the 1 μH inductors.

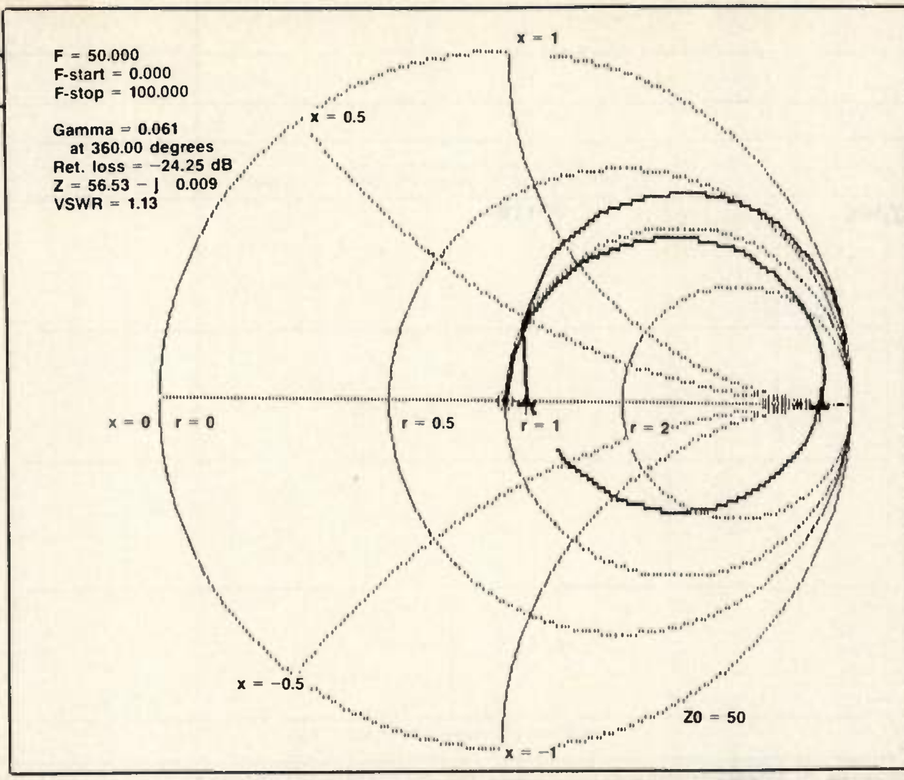


Figure 5. Input impedance versus frequency for the 50 MHz oscillator.

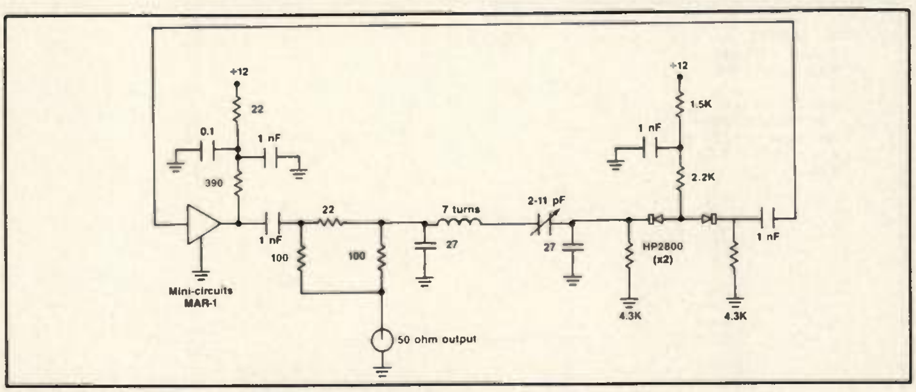


Figure 6. The 250 MHz MIC oscillator.

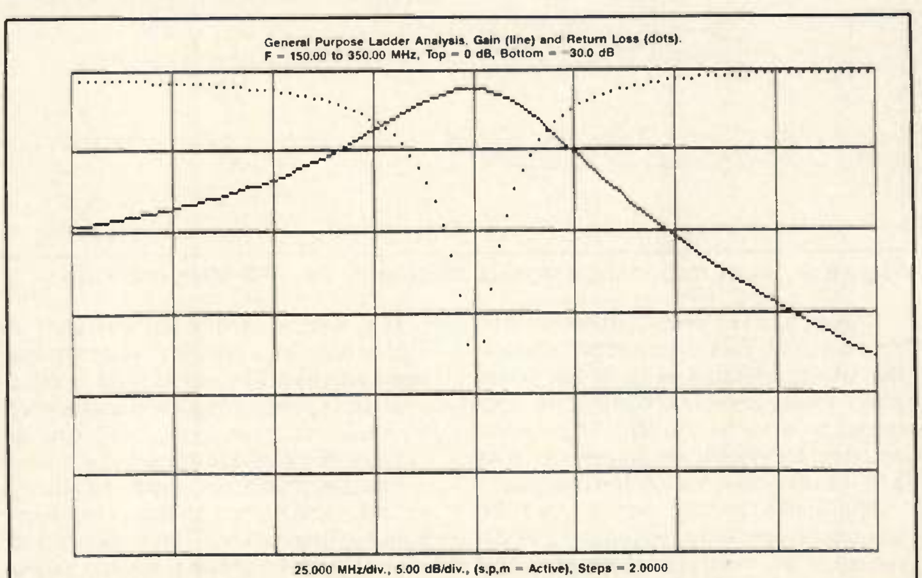


Figure 7. Resonator amplitude versus frequency for 250 MHz oscillator.

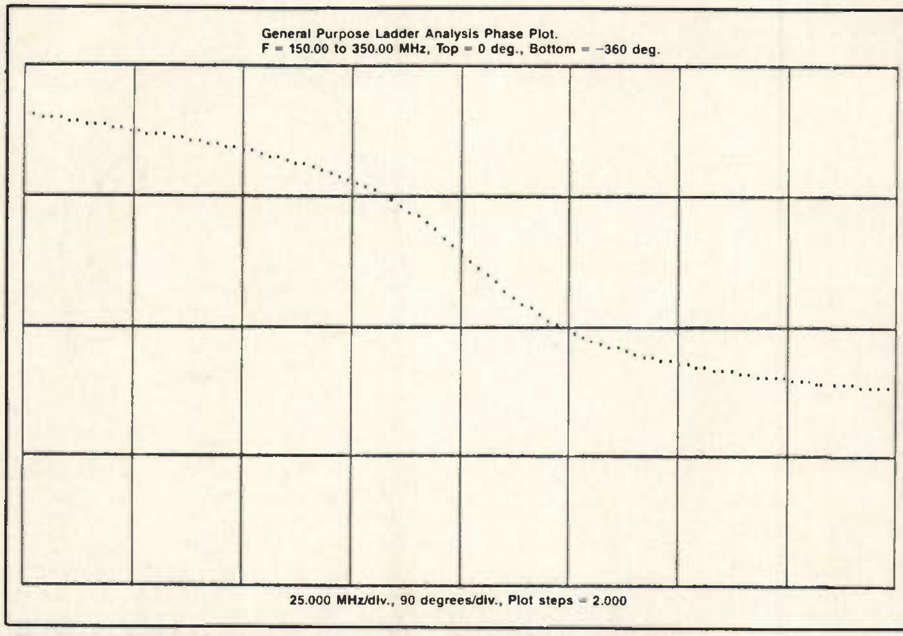


Figure 8. Plot of phase response for 250 MHz oscillator.

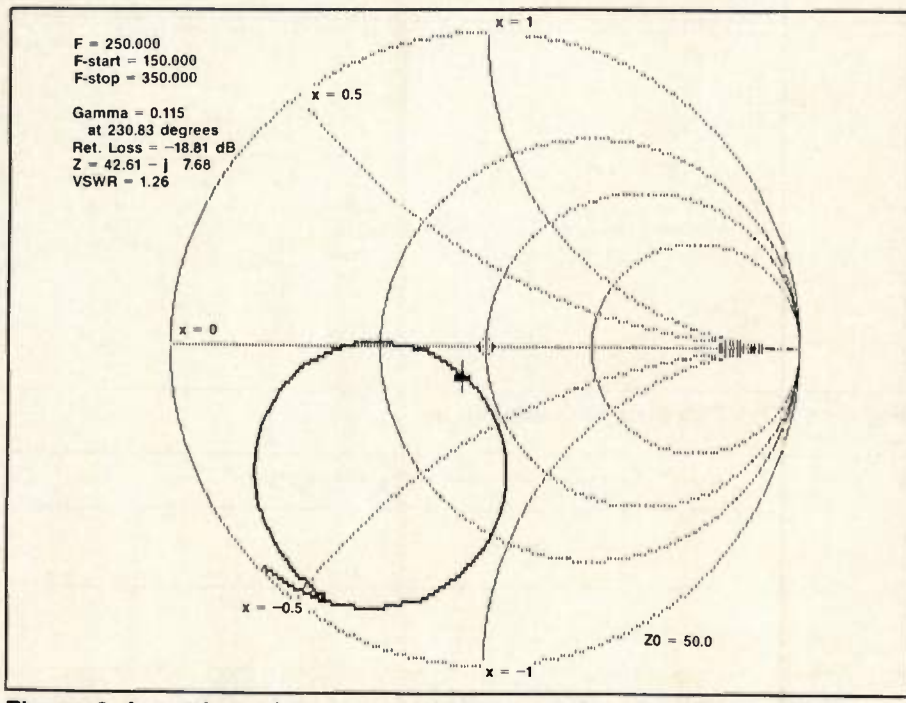


Figure 9. Input impedance versus frequency for 250 MHz oscillator.

This circuit had lower than normal harmonic output. This is a result of the amplifier which operates at fairly low linear power levels, with nonlinear limiting introduced by external circuitry. The second and third harmonics were measured at 44 and 48 dB below the 50 MHz output.

The series limiting mechanism used will cause operating resonator Q to decrease as resonator power increases. This is opposite to the effect desired for low phase noise.

The second oscillator is shown in Figure 6 where a different resonator has been adopted. The inductor used an air core with seven turns of #26 wire formed on a 6-32 machine screw, producing an inductance of about 75 nH. Calculated amplitude, phase, and input impedance for this resonator are shown in Figures 7, 8 and 9, respectively. The power divider was altered to provide slightly higher energy available within the loop. This decreased the available output power from the circuit which measured at −10 dBm. This oscillator operated at 250 MHz, which was the upper limit of the spectrum analyzer used for evaluation, which prevented measurement of harmonic distortion.

The amplifier used in the oscillator of Figure 6 is the Mini-Circuits MAR-1, specified for a 50 ohm transducer gain of 13 dB with a 1 GHz bandwidth. S-parameter data was unavailable for the amplifier. DC resistance measurements suggested that the amplifier topology for the MAR-1 was similar to the Darlington circuit used by Avantek (1). Examination of Avantek data sheets showed a well confined phase angle for S21. The diode limiter was temporarily eliminated from the circuit. The frequency change was minimal, but the amplitude increased by nearly 10 dB.

The MIC is a practical device for the construction of simple oscillators. The circuits presented show low harmonic distortion and predictable output power. MIC oscillators offer an additional advantage: they are formed from a cascade of 50 ohm characteristic impedance circuits. As such, they are more easily measured than other more traditional circuits. rf

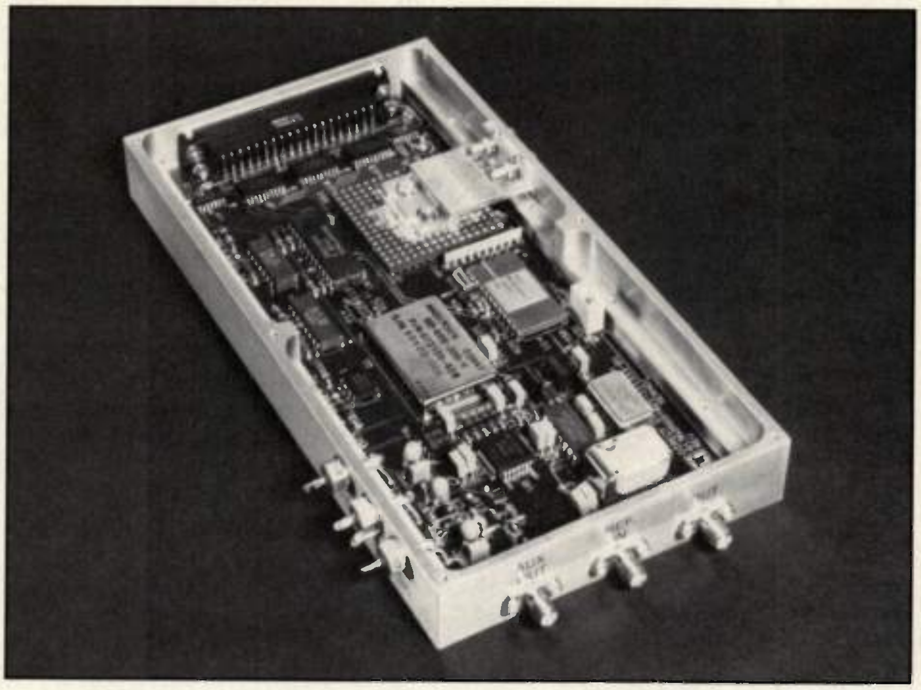
References

1. Snapp, Kukielka, and Osbrink, "Practical Silicon MMICs Challenge Hybrids," *Microwaves and RF*, Nov., 1982.
2. Hayward, "Introduction to Radio Frequency Design," Prentice-Hall, 1982. See chapter 7.
3. Boyles, "The Oscillator as a Reflection Amplifier: An Intuitive Approach to Oscillator Design," *Microwave Journal*, June, 1986.
4. Resonator analysis was performed with LADPAC, a program for the analysis and synthesis of ladder networks. The analysis methods are similar to those presented in *RF Design*, Sept./Oct., 1983.

About the Author

Wes Hayward is a senior design engineer at TriQuint Semiconductor, P.O. Box 4935, Beaverton, OR 97076, having recently transferred from Tektronix. Wes is a well-known author of books and articles on RF circuits and design techniques. He can be reached at (503) 627-3492.

130 MHz DDS from Stanford Telecommunications



The STEL-9272 is a digital frequency synthesizer based on the STEL-2172 ECL-NCO chip. It includes a 300 MHz clock generator which may be slaved to an external 5 MHz reference. The system also includes a parallel control interface and an anti-aliasing filter at the output of the digital to analog filter. It provides a signal in the 2 kHz to 130 MHz band with a resolution of just over 1 Hz. All spurious components are more than 37 dB below the carrier at all frequencies up to 130 MHz. The frequency control input data may be loaded as a parallel 28-bit word or as 4-bytes and frequency changes may be made as rapidly as every 125 ns.

In the byte load mode, the frequency selection data is loaded 1 byte at a time by means of the Load Enable lines. The bytes can be loaded in any sequence, and it is not necessary to re-load all the bytes when some of the frequency selection bits remain unchanged. In the parallel load mode, the frequency selection data is presented as a 28-bit parallel word and is latched in the system. The input buffer latches are transparent in this mode.

The system is packaged in an alloy casting with a mounting flange on the bottom surface. The control input connections are made through a 37-pin subminiature "D" type connector on one of the end faces, and the reference input and signal and auxiliary output connections are made through SMA connectors on the other end face. Power connections are made by feedthrough capacitors on one of the sides of the enclosure. In single quantities, the synthesizer is priced at $9750. **Stanford Telecommunications, Inc., Santa Clara, CA. Please circle INFO/CARD #204.**

Analogic Introduces Modular MRI Amplifiers

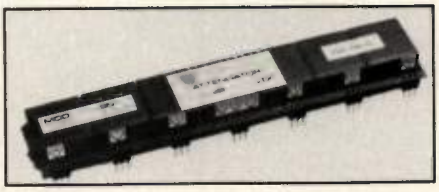
The AN80 Series provides single chassis pulsed power from 100 W to 1000 W and is field expandable from 2 kW to 4 kW with additional chassis. These amplifiers can supply RF power over a range of 5 to 90 MHz. The amplifier modules employ high power FETs in a configuration that combines Class A linear performance with Class B power efficiency. This is accomplished with high speed bias gating of the FETs. For high bias AB operation the gating is turned ON and for blanking the gating is turned OFF. The ON amplifier can produce low distortion RF with high envelope linearity.

The AN80 Series includes a peak RF power meter, an interface board and rear power-monitoring outputs. Transmission line transformers are utilized in the design since they pass relatively high frequencies and power levels with less dissipation than conventional RF transformers. The transformers are used as baluns and for impedance matching, splitting and combining. The two-to-one combiner transformers for the push-pull output stages also serve to reduce transistor distortion. The amplifier may be remotely controlled and monitored using the rear-panel interface board for communication between an external system and internal control board.

The specifications include a gain linearity of ±5 percent over a 40 dB dynamic range with gain flatness of ±2 dB from 5 MHz to 90 MHz. The typical third order intermodulation spurious response is −30 dBc and the typical noise figure is 12 dB. **Analogic Corporation, Peabody, MA. INFO/CARD #203.**

Latching Programmable Attenuators

The Models 5L80P and 7L80P, PCB mountable, latching programmable attenuators operate over DC to 1000 MHz and attenuate to 80 dB in 1 dB steps. The Model 5L80P is the 50 ohm version and the Model 7L80P is the 75 ohm version. The structure consists of four cells with



1, 2, 4, and 8 dB values, plus three cells with 10, 20 and 40 dB values. It is two attenuators in one package. The binary 1-8 dB portion provides 0 to 10 dB in 1 dB steps and the 10-40 dB portion provides 0 to 70 dB in 10 dB. The price is $145 in single quantities. **Trilithic, Indianapolis, IN. INFO/CARD #202.**

EMI Shielding Gaskets

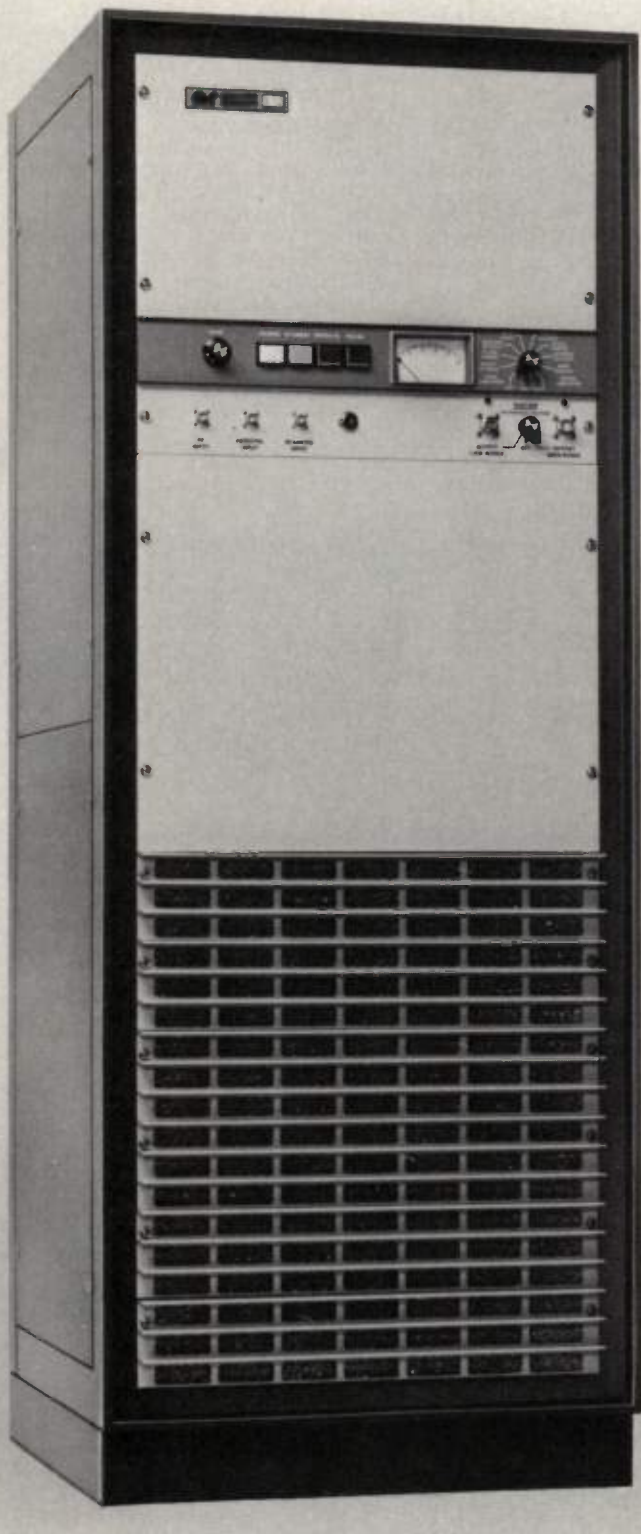
Schlegel® EMI shielding gaskets are made of compressible urethane foam encapsulated within conductive nylon ripstop fabric. The fabric is bonded to the foam and compresses easily. The average shielding effectiveness range is 69 dB when the frequency ranges from 1 kHz to 1 GHz. **Schlegel Corporation, Rochester, NY. INFO/CARD #201.**

Interference Reduction Filter

Cir-Q-Tel introduces a navigation satellite interference reduction filter. The high power low pass harmonic rejection filter was designed to replace the antenna coupler output insulator on AN/URA-38 and equivalent antenna couplers. The device will retrofit into antenna couplers in the 2-30 and 2-32 MHz communication

channels. It eliminates spurious emissions from antenna couplers which can overload critical satellite navigation systems operating in the 2-510 MHz range. **Cir-Q-Tel, Inc., Kensington, MD. INFO/CARD #200.**

Come where the power is.



Power amplifiers for every rf testing need. Power you can depend on. Power from one watt to ten kilowatts. And these power ratings are ultra-conservative: Whether you order a 10-watt or a 2000-watt amplifier, you can be certain of *at least* its rated output at every point in its frequency band.

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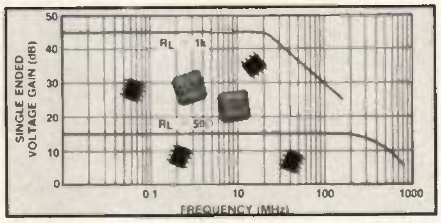
8292

400 MHz Amplifier

In applications where gain is a predominant factor, the SL6140 offers 45 dB gain into 1 k ohm load with a 25 MHz bandwidth. Where bandwidth factors dominate, operation over 400 MHz is possible into 50 ohm loads with 15 dB gain. Automatic gain is provided over a 75 dB range for



both cases. The cost is $3.50 in 1,000 piece quantities for the 8-pin small outline package. **Plessey Semiconductors, Irvine, CA. INFO/CARD #199.**

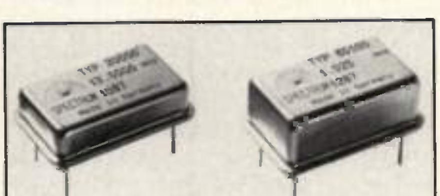
10 kW FM Transmitters

Thomson-CSF introduces the TH341 tetrode with associated cavity for 10 kW FM transmitters. It delivers about 17 dB gain at 80 percent efficiency. After initial

frequency tuning at installation, no further adjustments need to be made to the cavity. **Thomson Electron Tubes and Devices Corp., Dover, NJ. INFO/CARD #198.**

TTL Crystal Oscillators

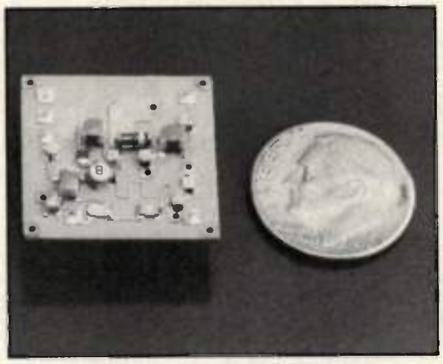
Spectrum Control unveils TTL crystal oscillators ranging from 0.2 to 100 MHz. The oscillators are built in hybrid thick film



technology and are hermetically sealed. **Spectrum Control, Inc., Erie, PA. Please circle INFO/CARD #197.**

Voltage Controlled Oscillators

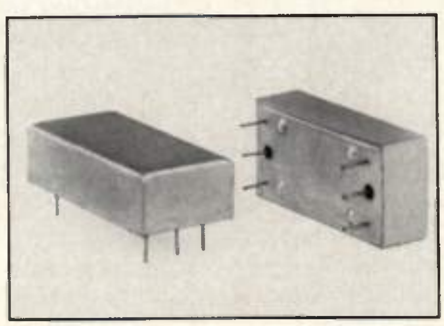
Z-Communications introduces the Model C-500 VCO which provides a 500 MHz tuning range from 950-1450 MHz, phase noise specification of −90 dBc at 25 kHz offset, 1 Hz bandwidth, harmonics at −30 dBc and power output of 11 dBm ±1 dB. Two versions with different voltage



requirements (+15 VDC and +12 VDC) are available. Samples are priced at $50 each to $15 in quantities of 10,000. **Z-Communications, Inc., Ft. Lauderdale, FL. INFO/CARD #196.**

Phase/Amplitude Matched Set Filters

PTI offers phase/amplitude matched set filters for military navigation and monopulsed radar guidance systems as well as spread spectrum communications applications. These sets are matched in filter pairs with a nominal frequency of 21.4 MHz and a 3 dB bandwidth of ±150 kHz minimum. The initial offset has a match-



ing phase of 10 degrees and amplitude of 1 dB. The spurious response is 60 dB minimum with an insertion loss of 10 dB maximum. It is priced at $650 each. **Piezo Technology, Inc., Orlando, FL. Please circle INFO/CARD #195.**

RF Power Amplifier System

The Model 4200 has a frequency range of 400-450 MHz with a minimum output power of 1 kW into 50 ohms. VSWR is 2:1 and gain is 60 dB with ±1 dB gain flatness. Spurious content is −60 dBc max. The 4200 is protected against infinite VSWR at rated power and will withstand input overdrive to +13 dBm. **American Microwave Technology, Inc., Fullerton, CA. INFO/CARD #194.**

SMT Inductors

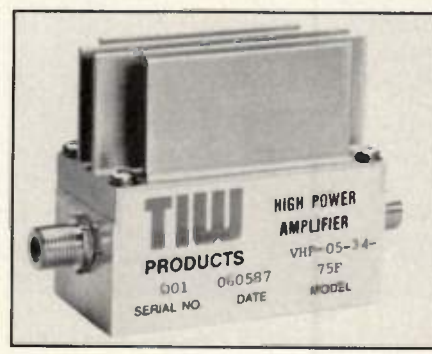
Coilcraft introduces surface mount inductors in a 1210 body size (.120 × .100 in.). They are available with inductance values from .01 to 10 uH and high Q's. They feature all ceramic construction and cost $0.19 each in 50,000 quantities.

Also being introduced is a SMD-A test fixture that is suited for testing tight tolerance, low value passive components at frequencies from 1 to 1000 MHz. The fixture has low stray capacitance and residual inductance. It accommodates body sizes 1008 to 1812 and connects

directly to any precision 7 mm test port. The fixture can be used with network analyzers, spectrum analyzers, and RF impedance analyzers. Price for the SMD-A test fixture is $425. A companion lab software package for testing surface mount inductors using a personal computer via IEEE 4888 bus is also available and costs $195. **Coilcraft, Cary, IL. INFO/CARD #193.**

High Power RF Amplifier

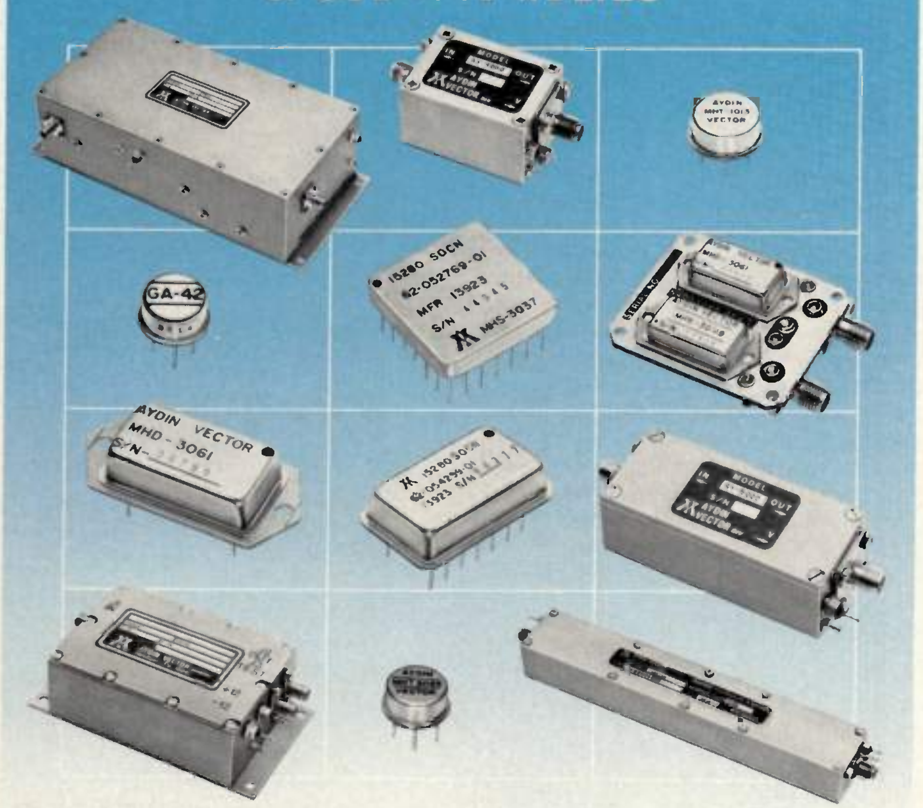
TIW Systems introduces the VHP-05 RF amplifier with a 40- 400 MHz bandwidth, 1 W continuous output and 34 dB nominal gain. The instrument amplifies several combined signals while generat-



ing few intermodulation products. The VHP-05 comes with 75 ohm input and output impedance and a choice of BNC, SMA or F type connectors. **TIW Systems, Inc., Sunnyvale, CA. INFO/CARD #192.**

Non-inverting PIN Drivers

The NX Series drivers provide steady-state output current with current spikes for fast PIN and NIP switching. Test points are provided to allow tailoring of output currents and spikes to particular applications. Internal current limiting assures that short-term accidental short-circuits to these test points will not damage the driver. The drivers have integral reverse bias protection and contain internal .01 uF bypass capacitors on both supply inputs. **Impellimax, Nashua, NH. INFO/CARD #191.**



Digital Telemetry Links

Neulink introduces a 928-960 MHz transmitter and receiver module for digital applications. The RFL-T9D transmitter and the RFL-R9D receiver are designed to transfer 700-2400 baud Manchester coded digital data. The transmitter has an adjustable power output of 1-2 W. Both receiver and transmitter feature 5 PPM stability. **Neulink, San Diego, CA. Please circle INFO/CARD #190.**

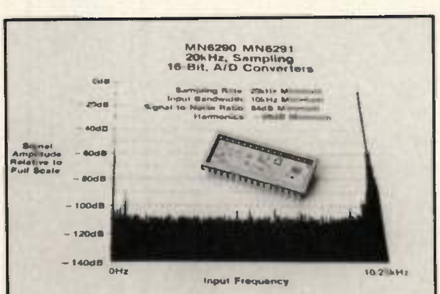
SMT Pi Filter

With typical insertion losses ranging from 40 dB at 100 MHz to 20 dB at 1 GHz, these pi-sectional filters deliver low residual self-inductance and high self-resonant frequency due to their use to a square dielectric in a feed-through con-

figuration. In addition, a low Q ferrite material provides a lossy series impedance. **Oxley, Inc., Branford, CT. Please circle INFO/CARD #189.**

A/D Converters

The MN6290 Series of low-distortion, 16-bit, sampling, A/D converters offer a combination of resolving power, conversion speed, low noise, and low harmonic distortion. They are packaged in 32-pin, double wide DIPs and have internal track-



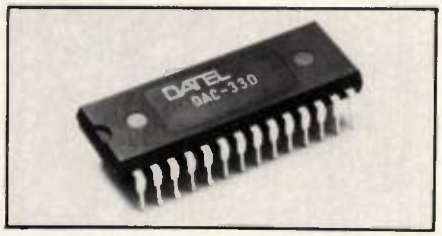
hold amplifiers that enable them to sample and digitize 10 kHz full scale input signals at rates up to 20 kHz. The signal to noise ratio is 84 dB. **Micro Networks, Worcester, MA. INFO/CARD #188.**

UHF Transmitter Combiners

The Antenna Specialists Co. introduces a series of transmitter combiners for applications in the 450 to 470 MHz range and features stripline technology. Models are available to combine either four or six transmitters. **The Antenna Specialists Co., Cleveland, OH. INFO/CARD #187.**

10-Bit DAC

The DAC-330 is a monolithic, 10-bit, 100 MHz DAC with a 14 MHz bandwidth multiplying capability. The device's digital inputs are ECL compatible. The voltage output range is 0 to −1 V, with a settling time



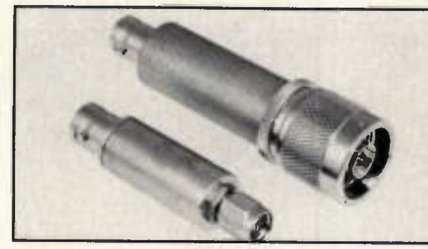
of 4.7 ns. The product is available in single quantities at $112 each. **Datel, Mansfield, MA. INFO/CARD #186.**

GaAs MMIC Switch

A family of MMIC switches and switch arrays based on GaAs MESFET technology is available from Plessey 3-5. The P35-4220 SPDT switch is for applications from 10 MHz to 6 GHz, with a minimum isolation of 30 dB, maximum solation loss of 1.9 dB at 6 GHz and a typical switching speed of 2 ns. The P35-4222 and P35-4223 SPDT switches for 10 MHz to 3 GHz applications and has a typical insertion loss of 1.2 dB with 35 dB minimum isolation at maximum frequency. The P35-4240 and P35-4241 switch arrays are for digital switching applications. **Plessey Three-Five Group, San Diego, CA. INFO/CARD #185.**

Broadband Coaxial Detector

The ACSP-2587NZ is a broadband coaxial Schottky diode detector that covers 500 MHz to 2 GHz with a k-factor



of 2000 mV/mW and response of ±0.2 dB over the band. The detector is priced at $81. **Advanced Control Components, Clinton, NJ. INFO/CARD #183.**

Aperture Coupled Bandpass Filter

Decibel Products offers the DB4200 6-cavity aperture coupled bandpass filter for transmitters or receivers. The unit operates from 806 to 960 MHz and accommodates 15 MHz bandwidth SMR trunked systems or 20 MHz cellular or international applications. Insertion loss is 1.0 dB at 15 MHz bandwidth, 0.6 dB at 20 MHz bandwidth and maximum VSWR is 1.25:1. **Decibel Products, Inc., Dallas, TX. INFO/CARD #184.**

Power MOSFETs

APT introduces the APT4530AN and the APT5030AN power MOSFETs (Power MOS IV). The APT4530AN features a V_{dss} of 450 V, V_{gs} of ±30 V, I_d of 16 A and I_{dm} of 64 A. The APT5030AN features the same specifications with a higher V_{dss} of 500 V. In quantities of 1000, the 5030 is $18.23 and the 4530 is $14.58. **Advanced Power Technology, Bend, OR. Please circle INFO/CARD #180.**

VDT Radiation Meter

This instrument features RMS detection and auto ranging for measurements of non-ionizing radiation from computer video display terminals. It measures both E and H fields using a single sensing

head with switch selectable elements. The sensitivity range for the E field is from 1 V/m to 2000 V/m and 1 mA/m to 2 A/m for the H field. The HI-3600 is priced at $895. **Holaday Industries, Inc., Eden Prairie, MN. INFO/CARD #182.**

Bipolar Amplifiers

Two small signal bipolar amplifiers, the BFQ 70 and BFQ 71, are available from the Special Products Division of Siemens

Components. These NPN transistors are designed for use in low- noise amplifiers and oscillators up to 2 GHz. The are available in standard 12 mm tape and reel format for surface mounting. The NF for the BFQ 70 is 1.8 dB while the NF for the

BFQ 71 is 2.0 dB. Pricing for 100 units is $6.32 for BFQ 70 and $5.98 for BFQ 71. **Siemens Components, Inc., Iselin, NJ. INFO/CARD #181.**

GaAs ICs for Fiber Optics

Anadigics introduces several GaAs ICs for use in fiber optics applications from 50 Mb/s to 2.7 Gb/s. The devices include a 3 Gb/s transimpedance amplifier (ATA30010), 10 kHz to 3 GHz amplifiers (ADA25002/3), 0.5 ns decision circuit (ACP10010), and a 350 MHz unity gain stable op-amp (AOP3510). **Anadigics, Inc., Warren, NJ. INFO/CARD #179.**

SMT Chip Inductors

Sprague-Goodman Electronics introduces the Surfcoil® line of chip inductors. Designed for surface mount applications, the inductors are standard in carrier and reel packaging. Specifications include inductances from 0.22 to 1000 uH. **Sprague-Goodman Electronics, Inc., Garden City Park, NY. INFO/CARD #229.**

ECL Clock

Model CO-233KEQ provides a stable 100K sub-nanosecond complementary ECL logic output at any specified frequency in the 150 to 500 MHz range. It is set to within ±.001 percent of the specified frequency. Surface mount internal construction results in a package size of 1.5"× 1.5"× 0.5". **Vectron Laboratories, Inc., Norwalk, CT. INFO/CARD #228.**

Low Cost Generator/Counter

Model ETG0410 combines the function of a 400 MHz to 1000 MHz signal generator and frequency counter. Output power level is 0 dBm to –70 dBm (calibrated) and +5 dBm (uncalibrated). The integral frequency counter has front panel access giving –15 dBm sensitivity from 400 MHz to 1000 MHz. **ED Tech, Inc., San Carlos, CA. INFO/CARD #227.**

Wide Band Filter

The FHLT/7-350/300-7/50-28A/28A has a center frequency of 350 MHz and a 1 dB bandwidth of 300 MHz. The VSWR at center frequency is less than 1.35:1 and insertion loss is less than 0.6 dB. It rejects greater than 15 dB at 192.5 and 555 MHz. **Cir-Q-Tel, Inc., Kensington, MD. Please circle INFO/CARD #226.**

Linear Amplifier Module

Model 700 is a MOSFET class A linear amplifier module capable of 2 W of power from 1 to 1000 MHz. Its linear gain is 26 dB (33 dB for Model 700A). Model 700 is $980 while Model 700A is $1050. **Kalmus Engineering International, LTD., Woodinville, WA. INFO/CARD #225.**

Feedforward Amplifier

A linear feedforward amplifier covering from 20 MHz to 200 MHz is available from WI-COMM. The 3rd order and 2nd order intercept points are typically 58 dBm and 92 dBm respectively. Noise figure is 6.5 dB and VSWR is better than 1.5:1. **WI-COMM Electronics, Inc., Massena, NY. INFO/CARD #224.**

RF Transformers

Coilcraft introduces a line of wideband RF transformers. They cover from .005 to 600 MHz with impedance ratios from 1:1 to 4:1. They are offered in tapped or untapped configurations. Applications include impedance matching, voltage or current transformation, DC isolation, balanced/unbalanced mixing, matching, power splitting, coupling, and signal inversion. An untapped part costs $1.80 each in 10000 quantities. **Coilcraft, Cary, IL. INFO/CARD #223.**

ECL Oscillator

Q-Tech introduces the QT6E1 (14 pins) and QT41E1 (4 pins) ECL oscillators that are available in a frequency range of 40 to 110 MHz. The typical supply current is 35 mA. **Q-Tech Corporation, Los Angeles, CA. INFO/CARD #221.**

Microwave Ceramic Capacitor

DLI introduces a single layer microwave ceramic capacitor for microstrip applications. The top side of the capacitor comes with a recessed (gold) pad and the bottom is fully metalized. It is available in sizes from 0.010" × 0.010" to 0.050" × 0.050" in nine dielectric materials to cover the capacitance range from 0.1 to 560 pF. Thicknesses available include 0.004" and 0.006". Typical frequency range is 2 to 26 GHz. **Dielectric Laboratories, Inc., Cazenovia, NY. Please circle INFO/CARD #222.**

rf software

Digital Filter Design Package

Atlanta Signal Processors introduces the Digital Filter Design Package 2.12 (DFDP2). In addition to a full range of IIR (infinite impulse response) filter design functions, it enables the user to design FIR (finite impulse response) filters to magnitude response specifications that can only be defined as piecewise-linear continuous curves. The FIR design modules can also design highpass, lowpass, bandpass or multiband filters as well as Hilbert transformers, differentiators, and raised-cosine pulse shaping filters. The IIR design module designs Butterworth, Chebyshev Type I and II, and elliptic filters. The package runs on IBM PC, XT, AT and compatibles, and PS/2 machines. DFDP2 sells for $1195. **Atlanta Signal Processors, Inc., Atlanta, GA. INFO/CARD #218.**

Dielectric Resonator Circuit CAD

This program allows users to efficiently approximate design with dielectric resonators, and features built-in help screens, menu driven format and hard copy capability. It provides users with a fast method to approximate within 2 percent resonant frequency of a particular resonator configuration. **Murata Erie North America, Inc., Smyrna, GA. INFO/CARD #220.**

IC Design Tool Kit

Analog Design Tools introduces a set of design tools for the Analog Workbench™. It is tailored for the simulation and analysis needs of IC designers. The IC Design Tool Kit, along with the Analog Workbench, is used to analyze the linear characteristics of IC designs, including linear, mixer-signal and digital ICs. It includes fabrication process data entry, which ties circuit behavior to the fab process for accurate simulation. Fundamental process parameters like N+ sheet resistivity, base junction depth and width, oxide thickness, and conduction factor can be defined by the designer. The kit can also be used to predict chip manufacturing yields before committing the design to silicon, using Monte Carlo analysis techniques, circuit sensitivity evaluation, and worst-case analysis. **Analog Design Tools, Sunnyvale, CA. INFO/CARD #177.**

rf literature

Copper Gasket Report

A report entitled *Transfer Impedance for Beryllium Copper Gaskets* is being offered by Instrument Specialties Company. The report includes updated gasket transfer impedance graphs measured using the current SAE-ARP-1720 calibration method. It also contains information relative to the test procedure and data regarding the test fixture and methodology used to measure RF gaskets for transfer impedance. **Instrument Specialties Co., Inc., Delaware Water Gap, PA. INFO/CARD #216.**

Catalog of Electronic Books

This catalog lists electronic technology books currently available from Prentice Hall. The listing includes books on electronic math, AC and DC circuits, electronic devices and active circuits, op-amps, linear ICs, satellite communications, microwaves and lasers, and various other related titles. **Prentice Hall, Englewood Cliffs, NJ. INFO/CARD #215.**

Brochure on Software and Applications Courses

This brochure from Integrated Computer Systems describes 13 software and applications technical short courses. The courses cover topics in hands-on programming, software development methods, knowledge-based systems, software tools and applications, hands-on advanced programming in C, and hands-on expert system design & development. For each course, the brochure describes the subjects and applications covered, the hands-on activities, benefits, materials provided, authors, instructors, locations and price. **Integrated Computer Systems, Culver City, CA. INFO/CARD #213.**

Audio Product Guide
This product guide features devices for professional audio applications, electronic music applications, general purpose, and other devices. The devices listed include voltage controlled amplifiers, microphone preamplifiers, level detectors, transconductance amplifiers, VCOs, voltage controlled filters, envelope generators, and music voice systems. **Solid State Micro Technology for Music, Inc., Santa Clara, CA. Please circle INFO/CARD #214.**

IF/RF/Microwave Catalog
Penstock announces the introduction of a catalog listing all companies and products that it distributes. Over 700 pages covering 11 companies, 31 product lines, and thousands of products types are included. **Penstock, Inc., Los Altos, CA. Please circle INFO/CARD #212.**

Catalog Features RF Equipment and Components
This catalog includes photos, specifications, features, outline drawings, and prices for Weinschel's line of RF and microwave components together with test and measurement instrumentation. Made for OEM, laboratory, military and commercial applications, Weinschel products cover from DC to 40 GHz and have a power handling capability ranging up to 500 W. **Weinschel Engineering, Gaithersburg, MD. INFO/CARD #206.**

Microstrip Application Note
Designing Microstrip Circuit with Low-Loss Cuflon Substrates covers microstrip design principles and discusses the advantages attained by using Polyflon's low loss Cuflon substrate in microstrip applications. Highlighted are equations used in microstrip design examples of how dimensions were determined for a microstrip power divider and directional coupler, and a listing of the effective dielectric constant in different thicknesses and widths. **Polyflon Company, New Rochelle, NY. Please circle INFO/CARD #205.**

Attenuator and Switch Catalog
JFW Industries has released a catalog containing product information on many of their attenuators and switches. New products in this catalog include plug-in, IP8T, and matrix switches. Specifications for the fixed, power, variable, and programmable attenuators are covered in detail. **JFW Industries, Inc., Indianapolis, IN. INFO/CARD #210.**

Frequency Allocations Chart
A USA frequency allocations chart that ranges from 10 kHz to 3 GHz is available from Motorola Semiconductor Products. The allocations shown are divided into nine categories: Standard Frequencies; Distress, Calling, and Search and Rescue; Amateur Radio; Broadcasting; Fixed and Mobile Services; Military, Aeronautical; Maritime; and Scientific and Space. A flat rolled copy can be obtained by circling the reader service number. **Motorola, Inc., Phoenix, AZ. INFO/CARD #175.**

Frequency Synthesizer Catalog
Frequency synthesizers produce precision frequencies governed by a high stability frequency standard. With fast, easy, remote programming, they are useful in advanced measurement or production systems and also serve as stand alone test equipment. Properties, specs, applications and prices of the PTS 040, PTS 160, PTS 250, and PTS 500 are given in the catalog. Data sheets on the PTS 120 and PTS Dual Output Synthesizer are also included. **Programmed Test Sources, Inc., Littleton, MA. INFO/CARD #208.**

Advertiser Index

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SAW Questions and Andersens.

Dispersive devices:

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QUESTION: What range of bandwidth, dispersion and center frequencies are available?

ANDERSEN: SAW dispersive devices are used for wide bandwidth applications (up to 500 MHz) with dispersions up to 100 μs. IMCON dispersive devices have narrower bandwidths but provide dispersions up to 600 μs (IMCON's have been cascaded to produce dispersions of 10 ms). Center frequencies of Andersen dispersive devices range from 1 MHz to 750 MHz.

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ANDERSEN: Just give us a call. We'll do everything we can to meet your needs. With our recent major staff increase, we offer one of the largest groups of SAW designers of any U.S. company.

If you simply want more information, send for our comprehensive handbook on Acoustic Signal Processing. Write to Andersen Laboratories, 1280 Blue Hills Avenue, Bloomfield, CT 06002. Or phone (203) 242-0761/ TWX 710 425 2390.

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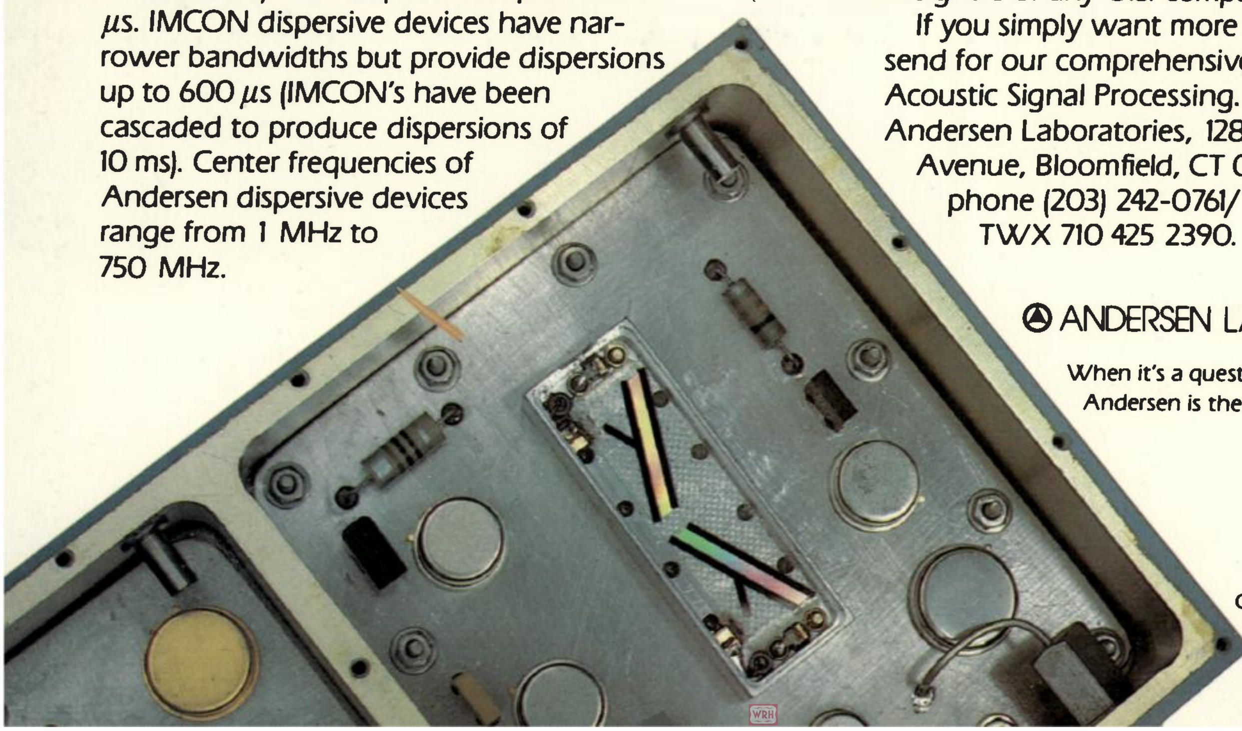
When it's a question of SAW, Andersen is the answer.



Compressive Receiver (actual size)

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