RF design

engineering principles and practices

May 1993

Cover Story
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INFO/CARD 116

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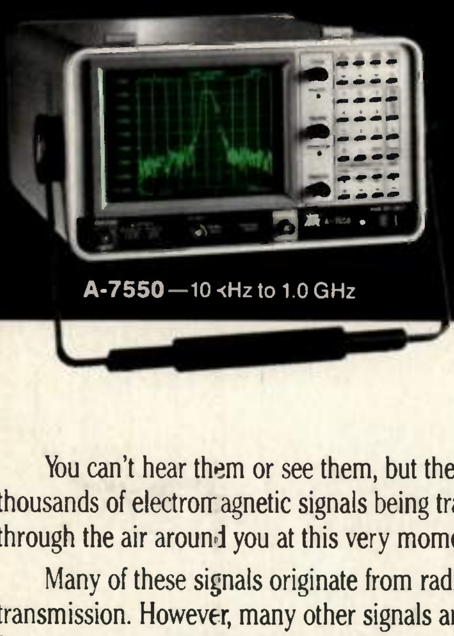

A-7550—10 kHz to 1.0 GHz

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You can't hear them or see them, but there are probably thousands of electromagnetic signals being transmitted through the air around you at this very moment.

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INFO/CARD 2

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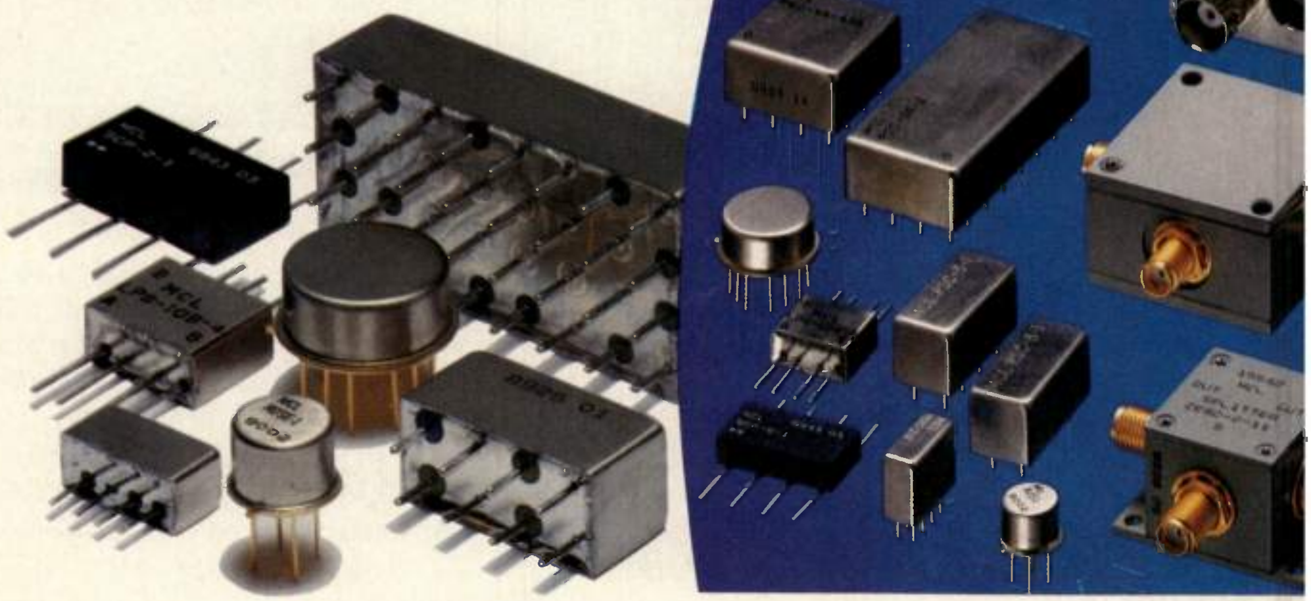

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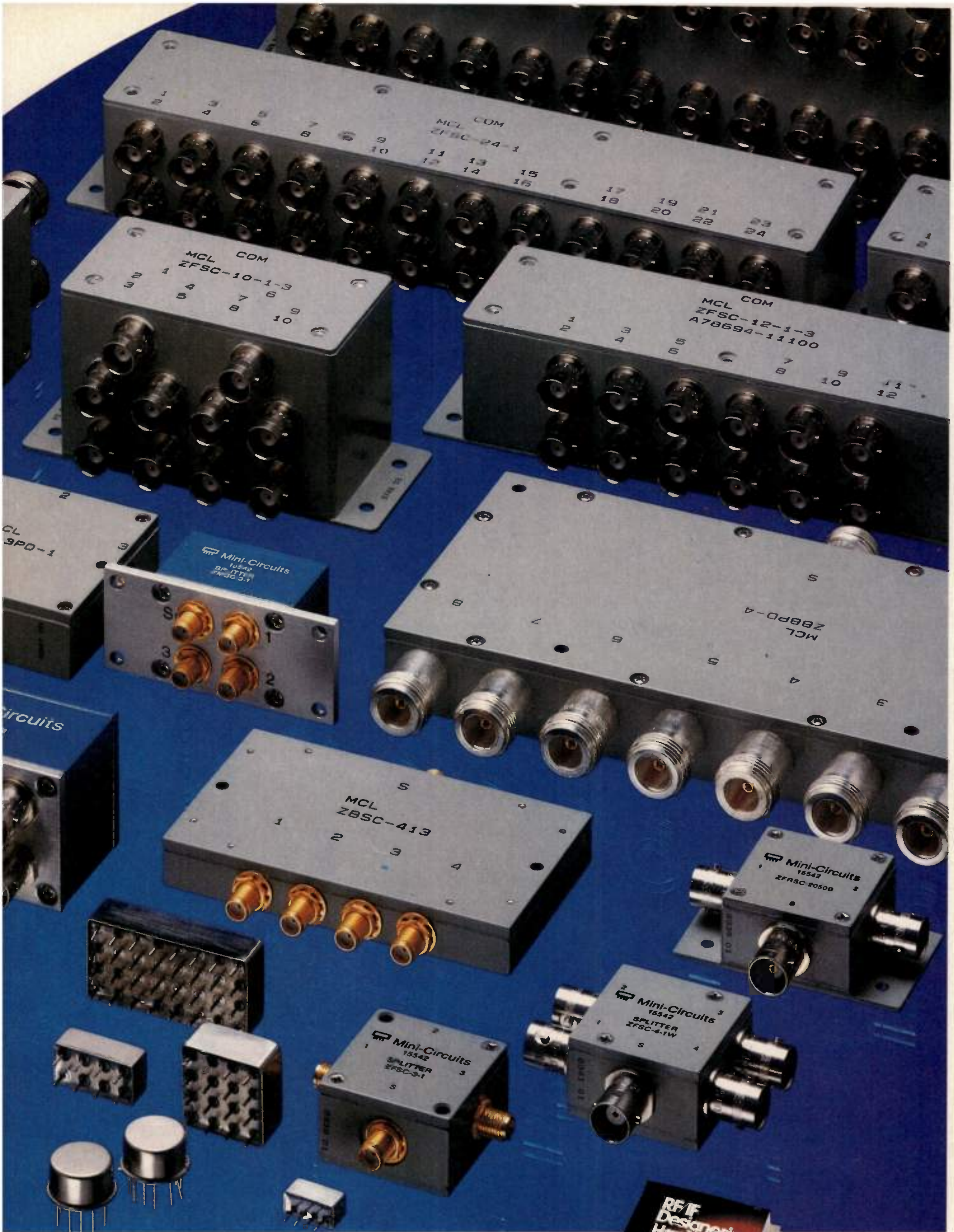

For detailed specs on all Mini-Circuits products refer to
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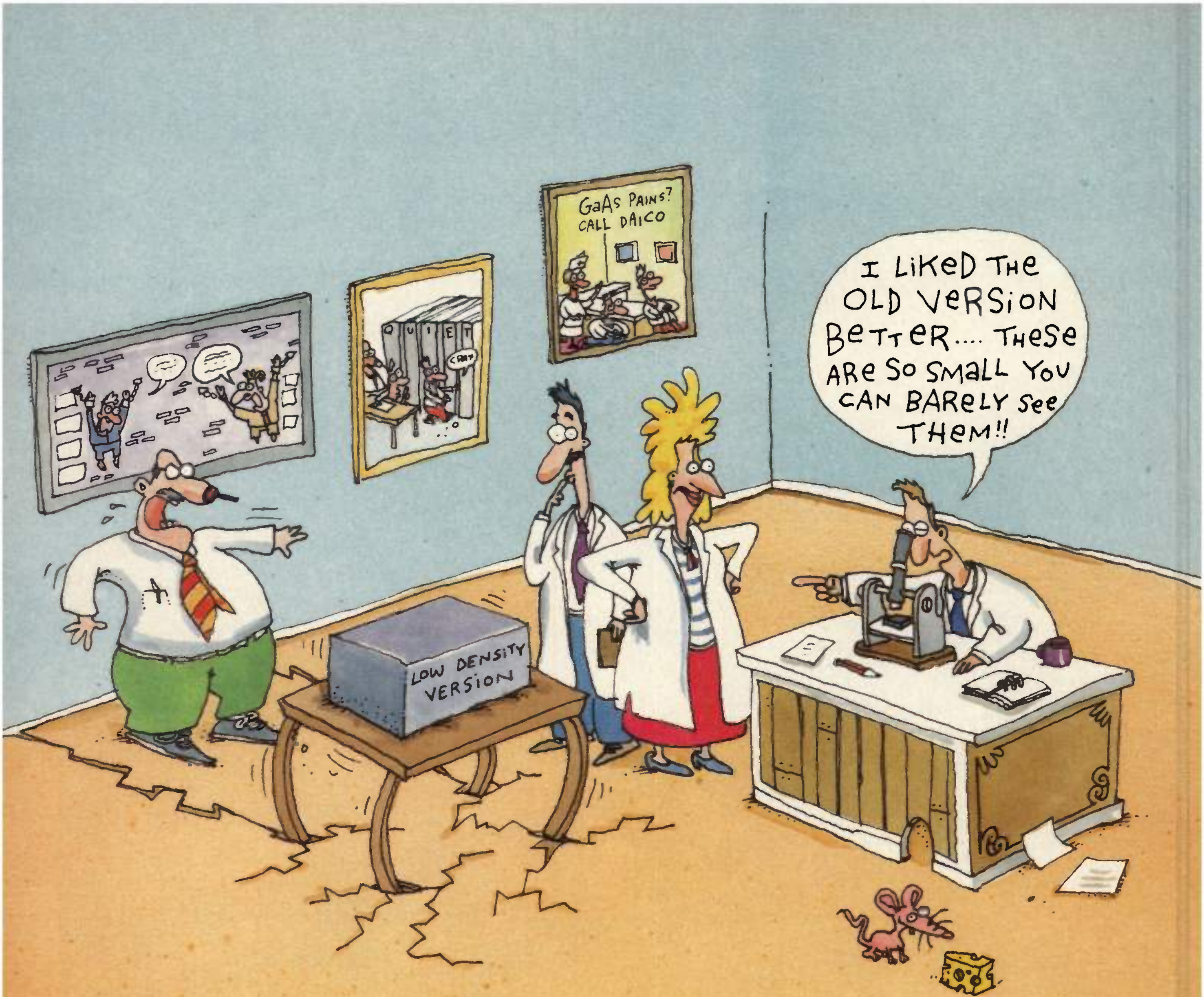

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RF design

May 1993

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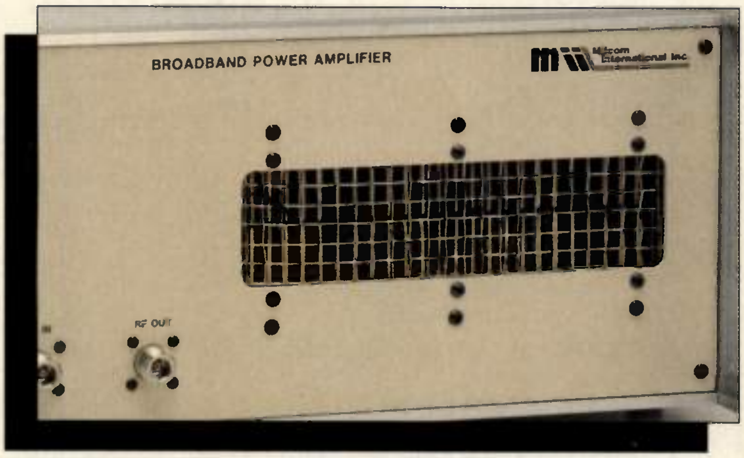

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Announcing A Special Addition to RF Expo East

By Gary A. Breed
Editor

We have held a lot of RF Expos! *RF Design* and Cardiff's trade show department recently completed the ninth (and again, very successful) RF Expo West, and the eighth RF Expo East is coming up. The value of these engineering forums has been confirmed over and over again by thousands of engineers who attend our courses and technical papers, and pursue their product search agendas in the exhibit hall.

RF Expo *PLUS*

For this October's RF Expo East, we are announcing our new RF Expo *PLUS* program — an extra technical track dedicated solely to Commercial Space Applications. This special conference covers an exciting major market for new RF products and technology. Nothing has been left out to make room for RF Expo *PLUS*, either. It is an added attraction on top of all the activities we have at RF Expo. If you want to take part, see our Call for Papers on page 60.

Although RF Expo East has seen a bit smaller attendance than its west coast sibling, the attending engineers have proven to be more serious. A high percentage of these engineers have taken our short courses and listened to the technical papers. RF Expo *PLUS* has been created in response to this demand for information. Note that Commercial Space is only the first topic to be featured; future Expos will highlight other timely subjects in this special program.

So, if you are already planning to attend RF Expo East, it's going to be everything you expected. If you are working on a space application intended for the commercial marketplace, it's going to be even more — you'd better not miss it!

Coming Events

RF Design will be on the road a lot in June. First, the Frequency Control Symposium is June 2-4 in Salt Lake City. Technical Editor Andy Kellett will be there, catching up on the latest developments in the technology of precision timekeeping.

The following week, June 10 and 11, is the Third Virginia Tech Conference on Wireless Personal Communications. In 1991, this conference was the first of its kind, covering engineering for new wireless communications systems. Dr. Ted Rappaport's Mobile and Portable Radio Research Group has pioneered educational and research efforts in this important area of RF technology. I'll see many of you in Blacksburg.

The *RF Design* staff will be in Atlanta for the MTT Symposium the week of June 14. This is a major gathering of engineers who work at microwave frequencies and the upper range of RF. If you are attending MTT-S, look us up at our exhibit booth.

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Bolstered by the worldwide strength and long-term stability of Hewlett-Packard, Avantek is now far more than a defense technology leader. We're a world leader, prepared to do what it takes to help you win. As partners, HP and Avantek are committed to serving the defense industry in a whole new way. See our message next month for more details. You'll find there's strength in numbers.

Look for our booth (#938) at the MTT-S show in June.

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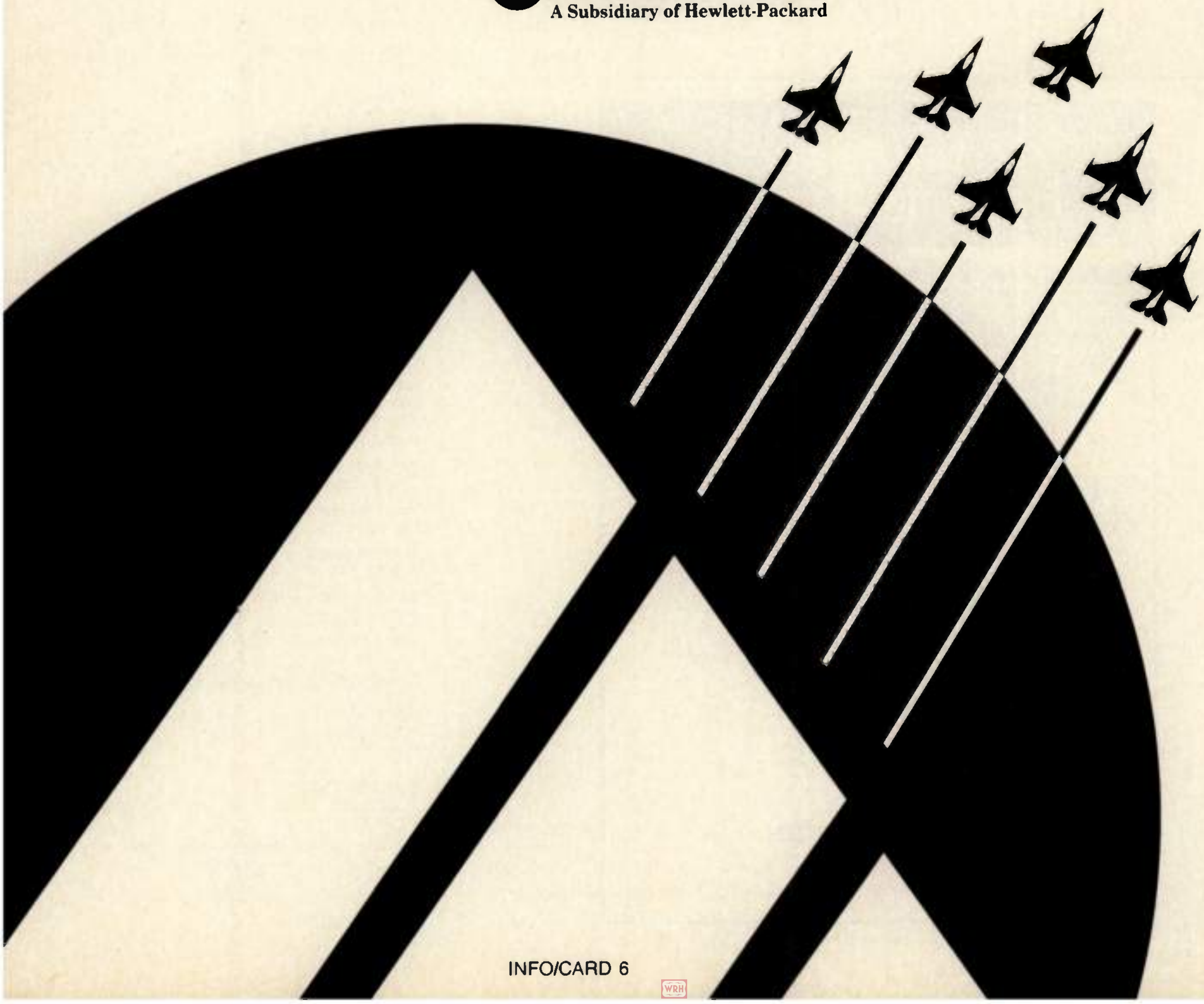

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INFO/CARD 6

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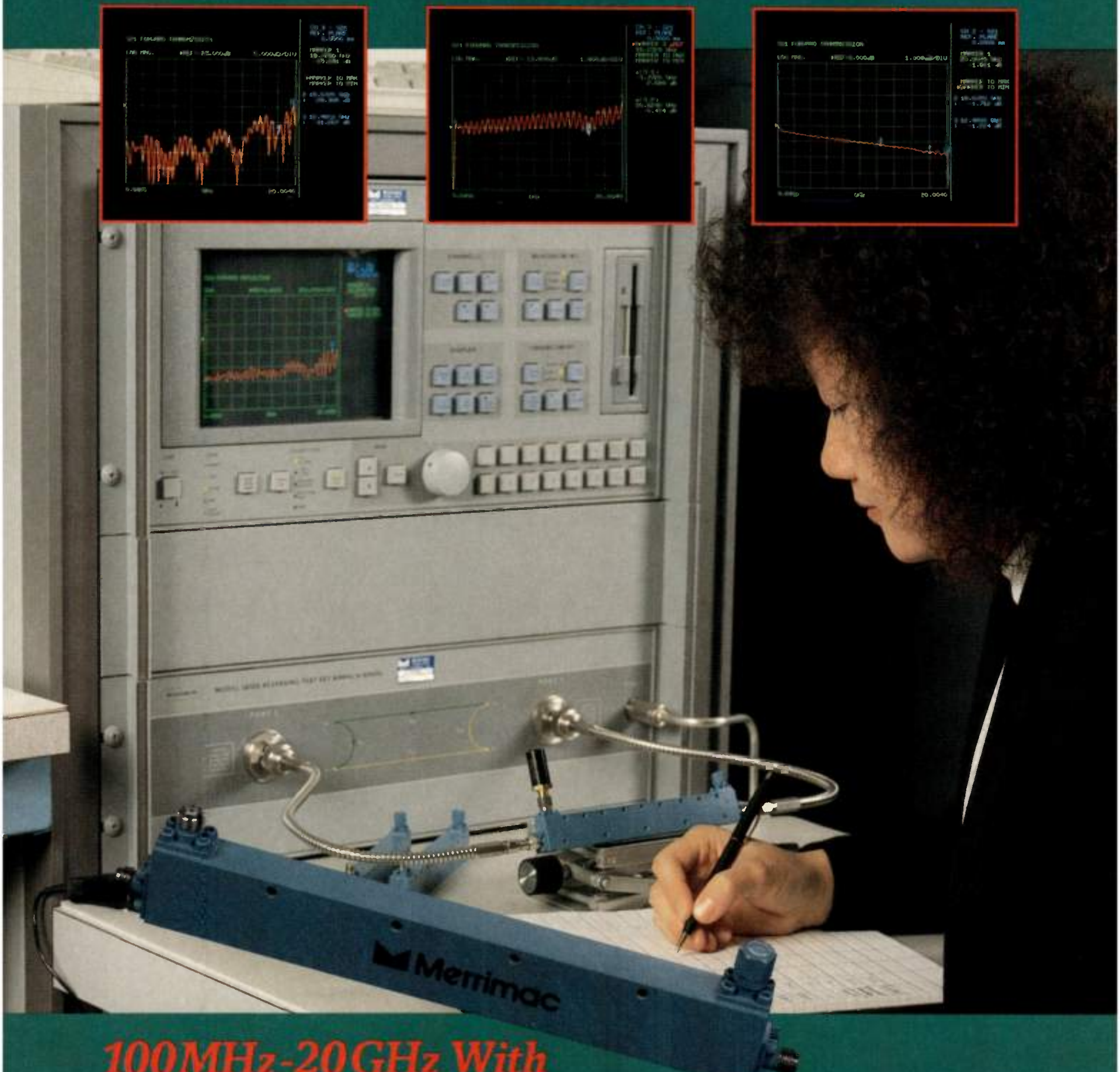

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GIGA-TRONICS 8540 UNIVERSAL POWER METER

Incredibly Fast and Accurate CW *and* Peak Power Measurements At A Truly Incredible Price.

UNIVERSAL POWER MEASUREMENT

Incredible is credible when describing the 8540 Series of Universal Power Meters.

From Giga-tronics, the new power in power meters.

For the very first time, you can make CW *and* peak power

The two-line display also lets you set the desired resolution and select either Lin or Log readout for each line.

measurements quickly and accurately with a single meter—a Universal Power Meter.

And all for about the same price you'd pay for the competitor's CW *only* power meter.

POWER MEASUREMENTS INSTANTLY

Imagine seeing display updates instantly: measurement speeds over the GPIB exceeding 200 readings per second and an exclusive Burst Mode capturing more than 2,000 readings in the same tick of a clock.

And because the 8540 Series uses diode sensors, you can measure all the way from −70 to +20 dBm with the same sensor, and without range changing delays.

If you're worried about having to write new code for your computer controlled testing, don't be: The 8540 Series uses the same GPIB command set as HP's 436A, 437B and 438A.

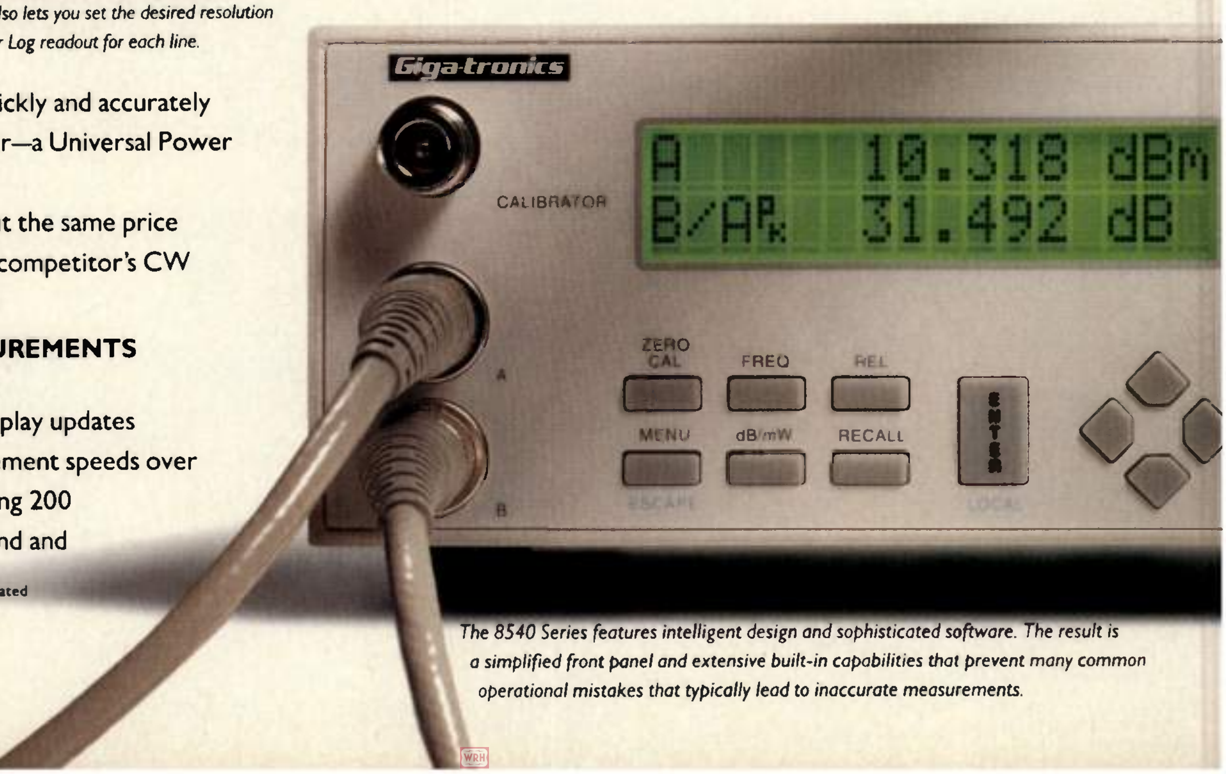
Think about what all this will do for your ATE productivity as well as for your company's bottom line.

FAST, EASY PEAK POWER MEASUREMENT

Now, an easy-to-use CW power meter can also measure pulsed RF signals with the simple addition of a peak power sensor.

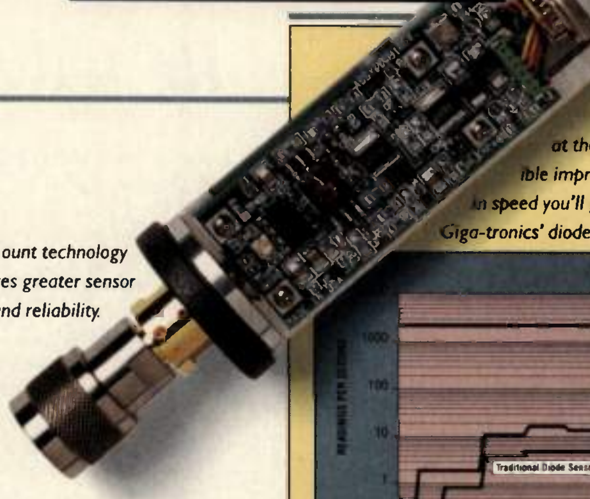
There are no time-consuming, unreliable duty cycle corrections, and you'll get the same accuracy and speed you'd get with a much-more-expensive dedicated peak power meter.

View the pulsed signal's amplitude profile on a scope and see the exact power measurement point on the pulse. Measure the overshoot. Measure the droop.



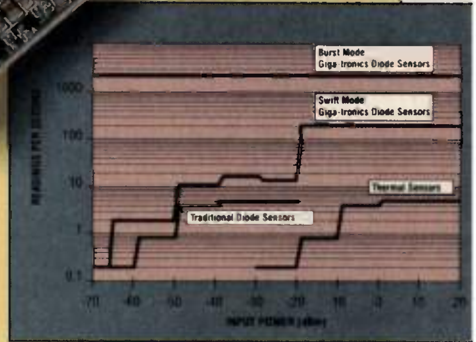
The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

The Secret Is The Sensors.



Surface Mount technology assures greater sensor accuracy and reliability.

Just look at the incredible improvement in speed you'll get with Giga-tronics' diode sensors.



You'll be confident of your peak power readings, and still have all the benefits of an incredibly fast CW power meter.

ONE OR <u>TRUE</u> TWO CHANNEL OPERATION

If a single-channel meter is what you need, the Model 8541 is the meter for you. But if you need two-channel capability, the Model 8542 lets you see readings from both channels *simultaneously.*

SIMPLE, INTELLIGENT OPERATION

The 8540 Series has only half as many controls as other power meters, but don't let that fool you. Intelligent design and sophisticated software give you easy access to extensive built-in capabilities.

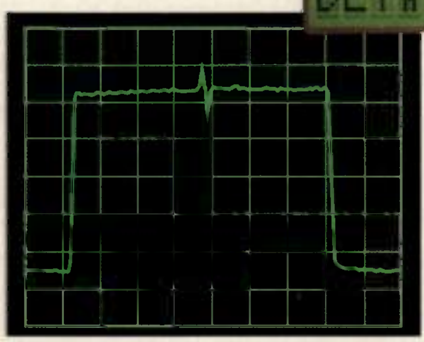
For example, you use the same key to zero and calibrate the power sensors. The meter automatically determines the function you want by detecting whether a sensor is connected to the calibrator.

Imagine all this power and performance. But why just imagine? Get the <u>truly</u> incredible Giga-tronics 8540 Series Universal Power Meter, and start measuring CW *and* peak power in a fraction of the time.

The Giga-tronics 8540 Series delivers incredible performance by taking full advantage of the speed and dynamic range of diode sensors.

What's more, Giga-tronics has solved the challenge that previously limited diode sensors to the "square law" region—below −20 dBm—by utilizing a built-in power sweep calibration system. So you get speed and a full 90 dB dynamic range without sacrificing accuracy.

A two-line back lit LCD display provides you more data in less time.



The peak sensor adds a marker on a monitor output for setting an exact measurement point on pulsed signals.

Call us toll-free at **1 800 726 GIGA.** Outside the U.S. call your local Giga-tronics representative, or call 001 408 734 5780. We'll send you more information or arrange for an incredible hands-on demonstration.

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RF letters

Letters should be addressed to: Editor, *RF Design*, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Published letters may be edited for length or clarity.

Expediency in the Far Field

Editor:

I enjoyed the article "Locating Power Line RF Interference" (February, 1992). As Operations and Maintenance Engineer for a large HF voice network, I often need to locate a source of HF noise around our station.

In most cases I can locate the offending appliance, fitting, insulator, etc. within 30 minutes using a simple travel-portable HF receiver. There is no RF gain control on my receiver, but adjusting the length of the telescopic whip adjusts the sensitivity. The same equipment does good service when preliminary site-surveying for customers.

A couple of years ago a nasty person cut some underground cable used to remotely control an HF transceiver. The newly laid cable was only shallow-covered for the night (when damaged), and we backfilled the trench the next day, fully unsuspecting.

Faced with excavating several hundred meters of trench to find the break, the first thought from university days was a TDR test- set. We don't have one. We set a signal generator to 100 MHz with FM tone modulation, injected it into the end of the cable, and walked along the cable route with the portable radio until the signal stopped. (I tried LF/HF first but found the cable radiated the signal too far to be useful.) It was a very distinct break in signal. Instead of digging up all of the cable (with risk of further damage), a one meter incision was made and was found to be spot-on. (Credit to colleague Manfred Kusterer for this idea.)

Simple equipment and techniques, but worth passing on. Keep up the good work with *RF Design*.

Nick Lock
Papua New Guinea

Reactance and Noise

Editor:

In the interesting article, ("Development of a Low Cost, Low Noise GPS Amplifier," February 1993, *RF Design*), the authors mention that the noise figure change is due to the resistance of the inductors used in the source circuit. While this is true for this illustration, in the generalized case one should also consider the possible additional change of the minimum noise figure, caused by the reactive part of the feedback circuit. In the low GHz region, due to the excellent low phase-shift characteristics of the GaAs FET, series inductive reactance does not significantly affect the minimum noise figure of the device, while it helps the stability. In a bipolar device, or a less capable FET where S21 phase characteristics are not as good, minimum noise figure and stability may be adversely affected (1).

Computer simulation of an active feedback stage reveals that by applying proper lossless (series and/or parallel) feedback, the minimum noise figure of a device may actually be reduced below what is specified on the data sheet. This may lead to erroneous conclusions that noise may be "tuned-out" by feedback. Unfortunately, under such conditions the gain is also reduced, so the noise measure of the circuit, which is a function of gain and noise figure, does not improve.

1. "Stability Considerations of a Low Noise Transistor Amplifier with Simultaneous Noise and Power Match," *IEEE MTT-S Int. Microwave Symposium*, May, 1975.

Les Besser
Besser Associates

Errata

In Ra'anan Sover's January 1993 article, "Switching Speed: Definition and Measurement," 34% should be replaced by 3µ% in Figure 3.

The caption for Figure 3 in April's tutorial, "A Basic Review of Feedback," should read "Feedback applied to an oscillator."

David Loker's article in the March issue, "An Introduction to RF/Microwave Filter Design," has the following corrections:

Equation 4 should be:

$$N \geq \frac{\cosh^{-1}\sqrt{\dfrac{10^{\frac{A}{10}}-1}{\varepsilon^2}}}{\cosh^{-1}(\omega_s/\omega_c)} \qquad (4)$$

Equations 17, 18 and 19 should read:

$$\text{series } L \rightarrow \text{parallel } LC \text{ (in series)} \qquad (17)$$

$$\text{where } L = \frac{g_i R \Delta\omega}{\omega_o^2},\ C = \frac{1}{g_i R \Delta\omega}$$

$$\text{shunt } C \rightarrow \text{series } LC \text{ (in shunt)} \qquad (18)$$

$$\text{where } L = \frac{R}{g_i \Delta\omega},\ C = \frac{g_i \Delta\omega}{R\omega_o^2}$$

where:

$$\omega \rightarrow \left[\frac{\omega_o}{\Delta\omega}\left(\frac{\omega}{\omega_o} - \frac{\omega_o}{\omega}\right)\right]^{-1} \qquad (19)$$

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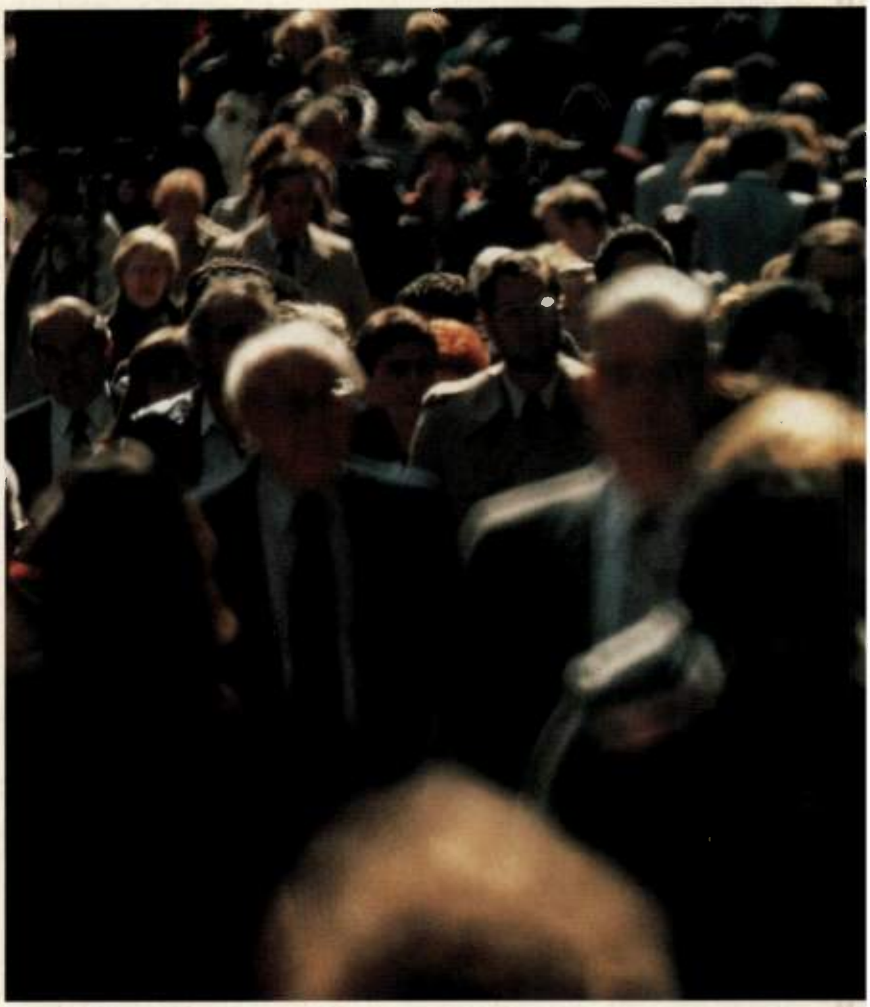

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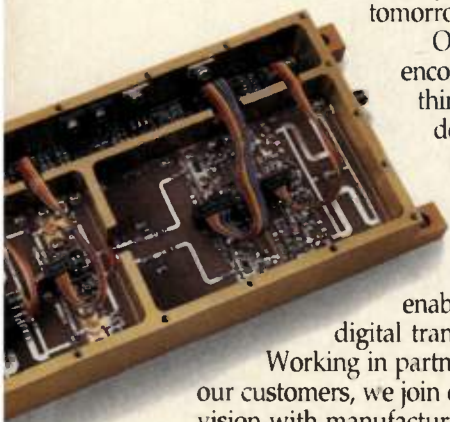

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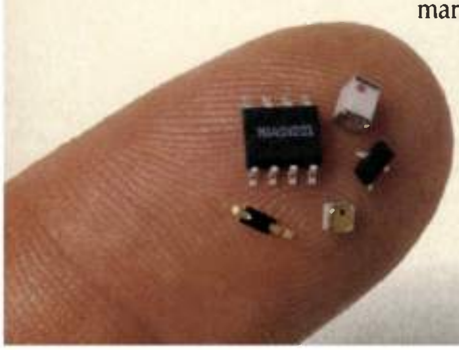

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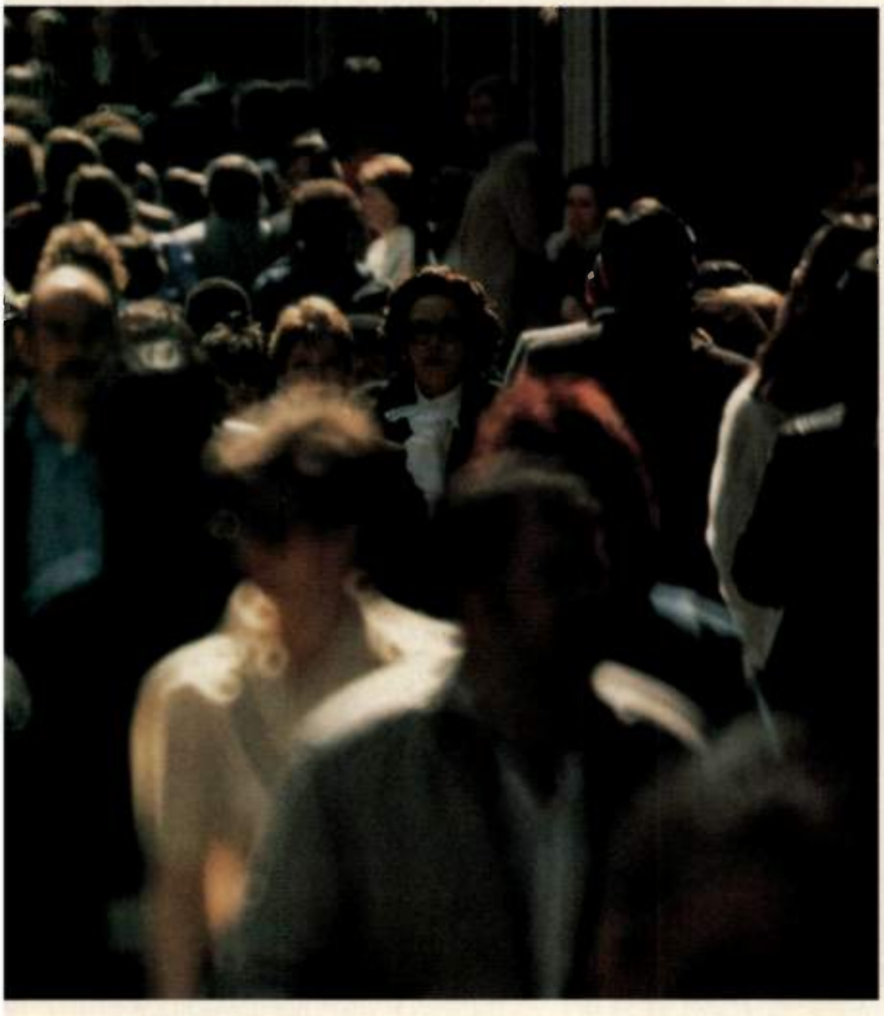

production of the more challenging and technologically advanced products the emerging digital and microcellular market requires.

Low-cost, high-power passive components.

Our new Linear Power Amplifier integrates a number of M/A-COM technologies including GaAs and Silicon MMICs (Monolithic Microwave Integrated Circuits), error correcting circuitry, and high-power RF transistors in a product that enables effective, high-quality digital power transmission. And our new cloverleaf pattern antenna is designed to solve the specific problems of crowded metropolitan microcells.

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M/A-COM

WRH

RF calendar

May

16-19 **1993 U.S. Conference on GaAs Manufacturing Technology**
Atlanta, GA
Information: Gary Troeger, Publicity Chair, 1993 GaAs MAN-TECH Conference. Tel: (714) 896-1692.

16-20 **ICA Comnet Dallas '93**
Dallas, TX
Information: World Expo Corporation, 111 Speen Street, P.O. Box 9107, Framingham, MA 01801. Tel: (800) 225-4698, or (508) 545-EXPO. Fax: (508) 872-8237.

18-20 **43rd IEEE Vehicular Technology Conference**
Secaucus, NJ
Information: Jesse E. Russell, Technical Chairman, AT&T Bell Laboratories, Whippany Road, Whippany, NJ 07981-0903. Tel: (201) 386-3314.

18-20 **IEEE Instrumentation and Measurement Technology Conference**
Irvine, CA
Information: Robert Myers, 3685 Motor Ave., Ste. 240, Los Angeles, CA 90034. Tel: (310) 287-1463. Fax: (310) 287-1851.

23-26 **ICC '93**
Geneva, Switzerland
Information: Peter Leuthold, Institut fur Kommunikationstechnik, ETH-Zentrum, CH-8092, Zurich, Switzerland. Tel: (41-1) 256-2788. Fax: (41-1) 262-0943.

June

2-4 **The 1993 IEEE Frequency Control Symposium**
Salt Lake City, UT
Information: Mr. Michael Mirarchi or Ms. Barbara McGivney, Synergistic Management Inc., 3100 Route 138, Wall Township, NJ 07719. Tel: (908) 280-2024.

9-11 **Wireless Symposium**
Blacksburg, VA
Information: Conference Registrar, Donaldson Brown Center for Continuing Education, Virginia Tech, Blacksburg, VA 24061-0104. Tel: (703) 231-5182. Fax: (703) 231-3746.

14-18 **IEEE MTT-S International Microwave Symposium**
Atlanta, GA
Information: John C. Hoover, Electromagnetic Sciences, Inc., Tel: (404) 263-9200 ext. 4245. Fax: (404) 263-9207.

18 **Automatic RF Techniques Group Conference**
Atlanta, GA
Information: Conference Chairman, Jonathan Schepps, David Sarnoff Research Center, MS 3-074, 201 Washington Road, Princeton, NJ 08540. Tel: (609) 734-2185. Fax: (609) 734-2034.

24-30 **Symposium on the Air-Sea Interface: Radio and Acoustic Sensing, Turbulence and Wave Dynamics**
Marseilles, France
Information: Dr. Michael SKAFEL, NWRI, CCIW, Box 5050, Burlington, Ontario, L7R 4A6, Canada. Fax: (416) 336-4989.

RF courses

Integrated Services Telecommunications Networking: High-Speed, Local, Metropolitan, and Wide-Area Networks
May 17-21, 1993, Los Angeles, CA
Hybrid Microcircuit and Multichip Module Packaging Technologies
May 24-26, 1993, Los Angeles, CA
Radiation Hardening of Electronic Systems
June 7-11, 1993, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (310) 825-1047. Fax: (310) 206-2815.

Modern Radar Technology: Monopulse Tracking Techniques and Other High-Performance Developments
May 24-28, 1993, Washington, DC
Analog/RF Fiber Optic Communications
May 26-28, 1993, Washington, DC
Anomalous Microwave and RF Propagation
June 2-3, 1993, Washington, DC
Electromagnetic Interference and Control
June 7-11, 1993, Washington, DC
Spread-Spectrum Communication Systems
June 7-11, 1993, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Navstar/GPS
June 2-4, 1993, Monterey, CA
Information: University Consortium for Continuing Education, Tel: (818) 995-6335. Fax: (818) 995-2932.

Eighth Vacation School on Data Communication and Networks
July 12-16, 1993, Birmingham, UK
Ninth Vacation School on Satellite Communication System
July 18-23, 1993, Guildford, UK
Information: The Institution of Electrical Engineers, Savoy Place, London WC2R 0BL, United Kingdom. Tel: 071-240 1871. Fax: 071-497 3633.

Computational Methods in Electromagnetics
June 7-10, 1993, Monterey, CA
Finite Element and Finite Difference Time Domain Methods for Solving Electromagnetic Engineering Problems
July 19-21, 1993, Worchester, MA
Information: Southeastern Center for Electrical Engineering Education, Kelly Brown - Registrar. Tel: (407) 892-6146. Fax: (407) 957-4535.

Aspects of Modern Military and Commercial Radar
June 14-18, 1993, United Kingdom
Error Correcting Codes and Trellis-Coded Modulation with Application to Communication Systems
June 7-10, 1993, Stockholm, Sweden
Analog CMOS Circuit Design for Signal Processing and Data Conversion
June 7-11, 1993, Stockholm, Sweden
Frequency-Time Signal Processing: Applications and Algorithms for High Resolution Spectral Analysis and Time Series Analysis
June 7-11, 1993, Stockholm, Sweden
Far-Field, Compact & Near-Field Antenna Measurement Techniques
June 7-11, 1993, United Kingdom
Information: CEI-Europe/Elsevier, Mrs. Tina Persson. Tel: (46) 122-175-70. Fax: (46) 122-143-47.

ELINT Interception, ELINT/EW Applications of Digital Signal Processing, Radar Vulnerability to Jamming
May 18-20, 1993, Syracuse, NY
ELINT/EW Data Bases
June 1-3, 1993, Washington, DC
Information: Research Associates of Syracuse, Tel: (315) 455-7157.

1993 High-Speed Digital Symposium
May 25, 1993, Dallas, TX
May 27, 1993, Austin, TX
June 2, 1993, Burlington, MA
June 4, 1993, Washington, DC
June 8, 1993, Boca Raton, FL
Information: Hewlett-Packard Company, Microwave Instruments Division (MID). Tel: (800) 765-9200.

Designing for EMC Seminar
May 17-18, 1993, St. Louis, MO
May 19-20, 1993, Chicago, IL
June 7-8, 1993, St. Paul, MN
EMC Diagnostics and Retrofitting Seminar
June 7-8, 1993, Rochester, NY
EMC Measurement and Regulation
May 24, 1993, Boston, MA
May 26, 1993, Albany, NY
May 28, 1993, Rochester, NY
June 14, 1993, Dayton, OH
June 16, 1993, Pittsburgh, PA
June 18, 1993, Detroit, MI
Information: Tektronix Inc. Tel: (800) 426-2200, (503) 629-4059. Fax: (503) 690-9307.

DSP Without Tears
June 7-9, 1993, Norcross, GA
June 16-18, 1993, San Jose, CA
June 23-25, 1993, Chicago, IL
July 26-28, 1993, Salt Lake City, UT
Advanced DSP With a Few Tears
June 10-11, 1993, Norcross, GA
Information: Z Domain Technologies, Inc., Tel: (800) 967-5034, (404) 664-6738. Fax: (404) 442-1210.

Applied High-Frequency Design Techniques I
June 21-25, 1993, Burlington, MA
Information: Hewlett-Packard HFDS Operation. Tel: (800) 472-5277.

Applied RF Design Techniques I
May 17-21, 1993, Rolling Meadow, IL
Information: Besser Associates. Tel: (415) 949-3300. Fax: (415) 949-4400.

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Model Number	Frequency (MHz)	Gain (Min.) (dB)	Gain Var. (Max.) (±dB)	Noise Figure (Max dB) Band Low	Mid	High	VSWR (Max.)	Output 1 dB Gain Comp. (Min., dBm)	DC Power (+V)	(@ mA)
AU-1310	.01 – 500	30	0.50	1.3	1.4	1.5	2:1	8	15	50
AM-1300	.01 – 1000	25	0.75	1.4	1.6	1.8	2:1	6	15	50
AU-1378*	1 – 300	17	0.50	1.9	1.9	1.9	2:1	–2	6	10
AU-1379*	1 – 500	13	0.50	2.2	2.3	2.4	2:1	–2	6	10
AU-2A-0150	1 – 500	30	0.50	1.3	1.4	1.5	2:1	8	15	50
AU-3A-0150	1 – 500	45	0.50	1.3	1.4	1.5	2:1	10	15	75
AM-2A-000110	1 – 1000	25	0.75	1.4	1.6	1.8	2:1	8	15	50
AM-3A-000110	1 – 1000	37	0.75	1.4	1.6	1.8	2:1	9	15	75
AU-1021	5 – 300	24	0.50	2.2	2.4	2.6	2:1	20	15	175
AU-1158	20 – 200	30	0.50	2.7	2.7	2.7	2:1	17	15	125
AMMIC-1318	100 – 2000	6	1.00	4.5	4.0	4.0	2:1	12	15	35
AMMIC-1348	100 – 2000	14	1.00	5.0	5.0	5.0	2:1	14	15	150
AM-2A-0510	500 – 1000	24	0.50	1.4	1.5	1.6	2:1	0	15	50
AM-3A-0510	500 – 1000	38	0.50	1.4	1.5	1.6	2:1	10	15	75
AM-3A-1020	1000 – 2000	30	0.50	1.8	2.1	2.4	2:1	10	15	75

*Designed for low current battery operation

POWER AMPLIFIERS

Model Number	Frequency (MHz)	Gain (Min.) (dB)	Gain Var. (Max.) (±dB)	Noise Figure Low	Mid	High	VSWR (Max.)	Output 1 dB Gain Comp. (Min., dBm)	DC Power (+V)	(@ mA)
AUP-1374	10 – 500	30	1.50	4.5	5.0	5.5	2:1	29	21	550
AUP-1383	20 – 300	35	1.50	2.4	2.5	2.6	2:1	29	21	650
AUP-1382	20 – 300	40	1.50	2.4	2.5	2.6	2:1	29	21	630
AMP-1380	10 – 1000	20	1.50	6.0	6.5	7.0	2:1	29	21	590
AMP-1381	20 – 1000	30	1.50	4.2	3.6	3.8	2:1	29	21	670
AMP-1389	10 – 1000	12	1.00	10.0	10.0	10.0	2:1	29	21	500

Other options available

For our amplifier catalog or for specific technical information, please contact Bill Pope at extension 282.

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INFO/CARD 17

WHITE NOISE NEWS

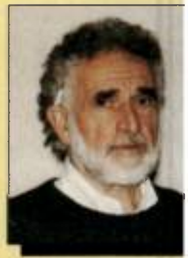
NEWS AND TECHNICAL INSIGHT FROM NOISE COM, INC.

ABOVE THE NOISE — BY GARY SIMONYAN

More than Simply New

With this issue of White Noise News, we introduce not just new products, but new capabilities as well.

Our noise source calibration service is simply the fastest, least costly method available. We do it in one day...with traceability to NIST.



Gary Simonyan

Our new UFX-BER Series instruments make bit-error rate testing easier than ever, and our NCT-6000 instrument requires only a spectrum analyzer to make noise figure measurements.

If you'd like fast information about any of these products, please call me at (201) 261-8797.

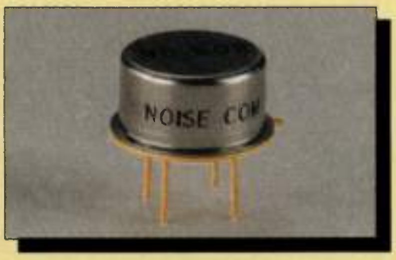
NOISE IN THE FIELD

Noise Sources Help Study the Ionosphere

An 8-channel receiver designed for SRI International and funded by Philips Laboratory is aiding the study of anisotropic irregularities of the electron density of the ionosphere. The anomalies can disturb or interrupt radar and radio communications.

The receiver uses a Noise Com NC502/15 as a built-in reference to provide the simultaneous amplitude and phase tracking necessary to estimate the irregularities. Between-channel tracking is possible to within 0.1 dB in amplitude and 0.1 degrees in phase. The built-in calibration includes timing of the sampling circuitry and linearity of the I/Q mixers and A/D converters.

INFO/CARD 117



FOR DIGITAL, SPREAD SPECTRUM

New Noise Instrument Simplifies BER Tests

Noise Com has introduced the UFX-BER Series precision C/N generating instruments that can set and maintain a precise ratio between a user-provided carrier and internally-generated noise. The instruments calibrate and display the ratio as either C/N, C/N_0, or E_b/N_0.

The UFX-BER is designed for use in IF back-to-back or RF loop-back testing of satellite, PCS, mobile, and CDMA spread spectrum communication systems. It is optionally capable of calibrating and displaying C/I for one or two user-provided interferer signals. This is useful for residual BER, adjacent channel, or CDMA testing.

Specifications (at IF or RF)

C/N Accuracy	+/-0.21 dB RSS +/-0.30 dB WCU

for -40 <C/N <+40 dB, -45 <carrier power <+15 dBm, and -45 <noise power <+3 dBm.

INFO/CARD 118

NEW FROM NOISE COM

Low-cost Instruments for Noise Figure Testing

Noise Com's NCT 6000 series low-cost instruments are designed exclusively for noise figure measurements. They'll provide precise measurements between 10 kHz and 18 GHz when used with a spectrum analyzer. Frequency resolution is the fine resolution bandwidth of the spectrum analyzer. The step attenuator accuracy is +/- 0.05 dB over any 1-dB range up to 1 GHz and +/- 0.25 dB over the full range.

The instruments contain an impedance-matching circuit, low-noise amplifier with 50 dB gain, precision 11-dB step attenuator with 0.1-dB resolution, power-conditioning supplies for 115 VAC, and the switches needed to perform noise figure measurements when using a spectrum analyzer.

INFO/CARD 119

NEW LITERATURE

Get Our Free New Catalog and Application Notes

We've just released two new application notes and a 24-page catalog that provide a wealth of information about noise sources.

Application Note 106 describes the use of white-noise diodes for the built-in test of balanced amplifiers. Application Note 112 covers the use of noise sources to calibrate the frequency response of spectrum analyzers. The catalog describes our complete line of broadband white-noise products that cover DC to 110 GHz, and includes information about noise calibration and applications.

All three are free, and you can get them by calling Noise Com at (201) 261-8797, or by circling the reader service number below.

INFO/CARD 120

THE WHITE NOISE ADVISER

What You Should Know About Calibration

Q: *How often should a noise source be calibrated?*
A: Noise sources are generally very stable, but it is industry practice that all noise sources and noise-based instruments be calibrated once every year, and even every 6 months for more critical applications.

Q: *Who can calibrate noise sources?*
A: Noise calibration requires special equipment such as metrology standards with calibration certificates issued by the U.S. National Institute of Standards and Technology (NIST). So they're not common items in the metrology lab. That's why Noise Com has been performing these services in compliance with MIL-STD-45662 for many large corporations as well as small businesses.

Q:: *How long does the calibration take?*
A: At Noise Com, we calibrate noise sources every day. All noise sources are calibrated and returned on the same working day as we receive them. So if you ship it to us for morning delivery, we'll send it back...calibrated...that afternoon.

Q: *What about NIST traceability?*
A: Noise Com maintains a metrology noise standard system that is calibrated directly at NIST, which provides the best accuracy. Noise Com also performs proficiency tests to periodically verify the accuracy of the calibration system.

Q: *What about my AIL, Eaton, HP, IMC, MDF, MSC, SGS-Thomson, Wavetek, or other noise sources?*
A: Noise Com calibrates all brands of noise sources and noise-generating instruments.

NOISE/COM

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RF Expo West Brings 3500 to San Jose

Attendance and activity at the ninth RF Expo West demonstrated the strength of the RF business. Development of new products for emerging applications, combined with adoption of new attitudes for cost-efficient products for existing applications, created an atmosphere of vitality and optimism.

A record number of exhibiting companies (190) gave engineers an opportunity to see a wide range of RF products. The only complaint heard about the exhibition was not having enough time to see everything on display. Typical of exhibitors' response to the show was this comment from Sharon Turcotte of Morrow Technologies: "The show has been real good for us."

The show was also real good for several companies with brand new products announced for the first time at RF Expo West. RF Monolithics introduced the HX1000 miniature transmitter for remote control, keyless entry and security applications. Sawtek brought a new hot-off-the-presses catalog and applications guide, Milcom International brought a new wideband amplifier, and Toko introduced a new line of chip inductors touted as the world's smallest.

In instruments, Marconi announced a handheld RF power meter with performance for either portable or lab use. For development engineers, another product of interest was Harris Corporation's major upgrade of its FASTRACK design software system for analog and mixed-signal ASICs. And these are just a few of the new developments seen by the attending RF engineers.

The new Introduction to RF Circuit Design classes were well-attended and well-received. Georgia Tech professors Dave Hertling and Bob Feeney provided a solid foundation for new engineers and a good review for more experienced engineers who attended this two-day course. Two additional courses, Filter and Matching Network Design and Oscillator Design Principles, again got rave reviews for their informational content and Randy Rhea's clear, well-paced instruction.

Finally, the technical papers were a primary attraction for RF Expo attendees. This year's conference featured an excellent series of papers on power amplifiers, plus timely presentations on cellular radio and personal communications. A wide variety of papers covered such key RF topics as miniaturization, frequency synthesis, body-worn antennas, ASIC development, computer modeling, and circuit analysis using the driving point impedance technique.

Call For Papers for RF Expo East and RF Expo Plus — RF Expo East, to be held October 19-21, 1993, in Tampa, Florida, has issued a call for papers. Papers are being accepted in the following areas: personal communications; data communications and DSP; specialized design techniques; new wireless applications; cellular and mobile radio; test, measurement and analysis; and essential RF circuits. In addition, this year's Expo will feature RF Expo Plus, a new addition focusing on

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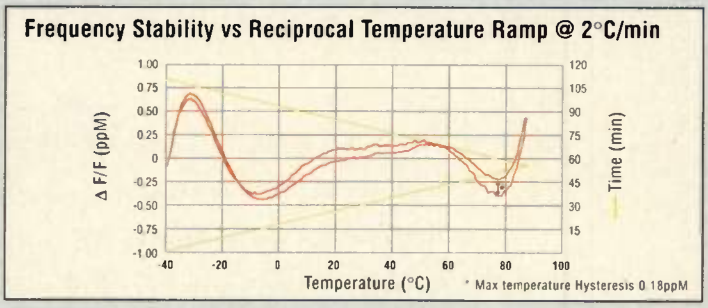

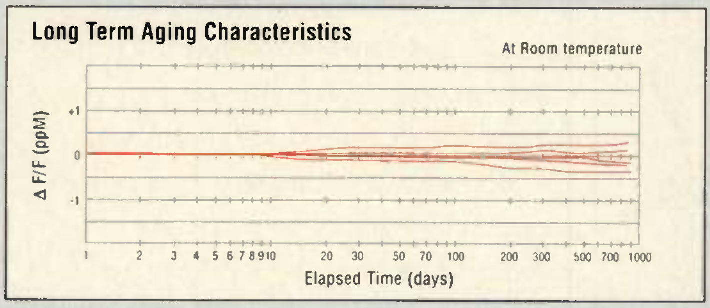

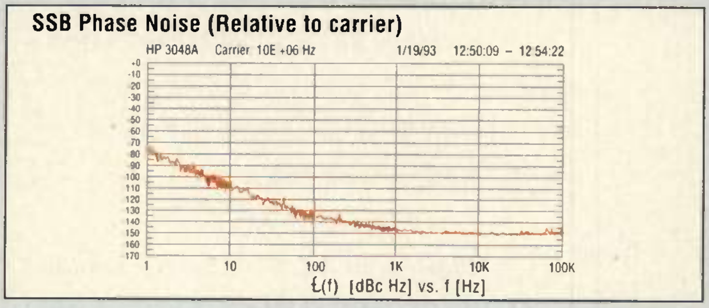

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INFO/CARD 124

Commercial Space Applications. Papers are also being accepted for this special track. Send a 50-word summary and brief outline by May 15 to RF Expo East, RF Design magazine, 6300 S. Syracuse Way, Suite 650, Englewood, CO 80111. Tel: (303) 220-0600, Fax: (303) 773-9716. Persons interested in moderating technical sessions are also needed and should reply to the above address.

Program Announced for 3rd Wireless Conference — The 3rd Virginia Tech Symposium on Wireless Personal Communications will be held June 9-11 on the university's Blacksburg, VA campus. Program highlights include sessions covering PCS, spread spectrum, propagation, and system modeling and simulation. These topics are addressed at the level of components, circuits and systems, from both theoretical and practical perspectives. An added feature is a Thursday evening panel session, with pioneers in cellular radio development discussing the evolution of that important wireless communications system. For more information, contact the Virginia Tech Center for Continuing Education at (703) 231-5182

Piezoelectric Devices Conference Call for Papers — The 15th Piezoelectric Devices Conference to be held September 21-23, 1993, at The Westin Crown Center, Kansas City, Missouri has issued a call for papers. Papers may be tutorial or application-oriented, dealing with recent progress in design, development, processing or manufacturing control in the following topics: properties of natural and cultured quartz, design of quartz resonators, oscillator and filter design, CAD/CAM, manufacturing and process control, measuring and test techniques, quality and reliability assurance, SAW devices and applications, and other piezoelectric devices and materials. Mail three copies of the abstract with your phone number as soon as possible to: Components Group, Electric Industries Association, 2001 Pennsylvania Ave., N.W., Washington D.C. 20006-1813.

Schlumberger and Motorola Form Joint Venture — Schlumberger Limited and Motorola Inc. recently announced that they have formed a joint venture to develop wireless electronics technology for remote and automated meter reading (RAMR). The joint venture, named Advance Meter Reading Technologies, is a design center that will develop wireless RAMR products and provide integrated solutions for water, gas, heat and electricity utility meters on a global basis. The joint venture will be equally owned by both companies with equal representation on its board. The joint venture is intended to develop the next generation of cost-effective utility meter reading technology to better serve this rapidly growing trend.

IVHS Consortium Formed — A consortium has been formed of five companies, M/A-COM, Inc., IVHS Technologies, Allstate Insurance Company, Eaton Corporation and AIL Systems, Inc., to focus on developing new products for the emerging Intelligent Vehicle Highway Systems (IVHS) Industry. Products for this industry include a range of automotive and vehicle sensors

INFO/CARD 22

such as collision avoidance systems, ground speed detectors, near object detection systems and others providing road-to-vehicle communications and vehicle-to-vehicle communications.

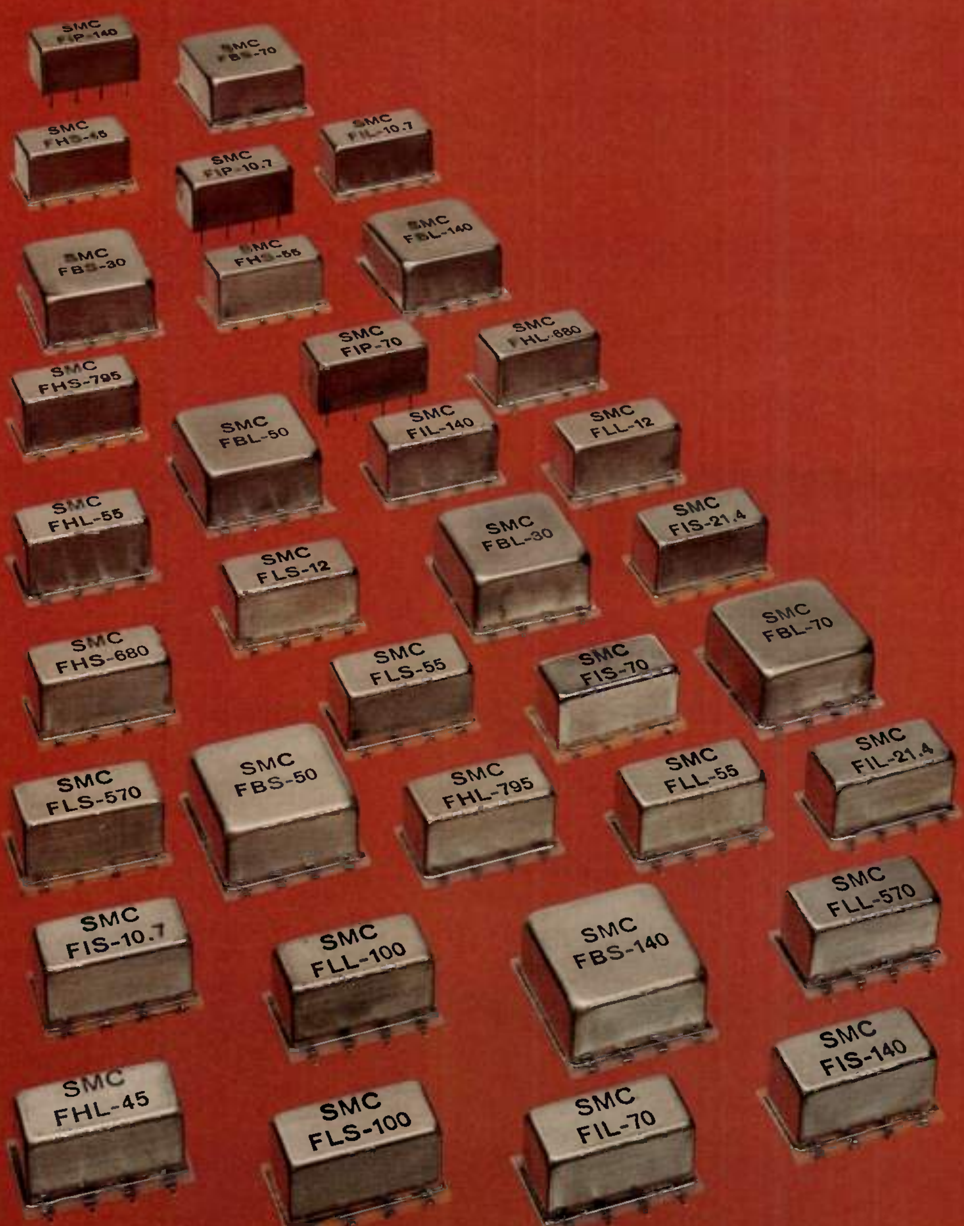
Watkins-Johnson Awarded Subsystems Contract — Watkins-Johnson Company recently announced that they have been awarded a contract valued in excess of $1 million by the Accelerator Systems Division of the Argonne National Laboratory. Under the contract, Watkins-Johnson will supply monopulse receiver subsystems for use in the beam-position-monitoring system for the Advanced Photon Source Program.

First Call for Papers — The IEEE GaAs IC Symposium, focusing on developments in integrated circuits using GaAs, InP, and other compound semiconductor and heterostructure devices, is now accepting papers in the following areas: materials and processes; CAD/CAM/CAT tools; device research and development; new device applications; high speed architecture; circuit design; circuit fabrication; circuit applications; automated testing; packaging technology; manufacturing technology; cost reduction methods; reliability and quality assurance; commercial products; and military and commercial systems. The symposium will be held October 10-13 in San Jose, California. Mail forty (40) copies of a one-page abstract and any supporting material by May 8, to: Donald C. D'Avanzo, Technical Program Chairman, 1993 IEEE GaAs IC Symposium, c/o Courtesy Associates, Attn: Cathy Coyle, Suite 300, 655 Fifteenth Street, N.W., Washington, DC 20005. Tel: (202) 347-5900. Fax: (202) 347-6109.

Contactless Smart Card Venture — Trendar Corporation and AT&T have announced that they are teaming up to provide a transaction station for truck stops based on AT&T's smart card technology. The contactless smart cards have microprocessors and memory chips laminated within their plastic shells. Developed at AT&T Bell Laboratories, their electronic memory can hold the equivalent of several pages of typewritten information, personalized to the card's user. Truck drivers will use the cards to fuel their vehicles and speed up the transaction process. The smart card will store such information as the truck number, purchase authorization limits, the driver's license number, the name of the fleet operator and any discount to which trucks in that fleet are entitled.

New Cruise Ship Direct Dial Phones — Two new systems being instituted by Norwegian Telecom will allow cruise ship passengers and anyone else at sea to call almost anywhere in the world via satellite. One is a new Inmarsat coast earth station, the other, a new credit card telephone service. The third Inmarsat earth station, to become operational in March, will extend Inmarsat coverage in North and South America, including the Miami-Caribbean region. The telephone service being launched at the same time, will enable passengers to call any telephone on land direct by pulling a card through the magnetic-strip reader on the telephone before dialing the desired number.

FILTERS



Synergy has introduced a selection of high performance, low cost standard catalog and custom highpass, lowpass and bandpass filters covering passbands within the DC to 2000 MHz range. These filters are readily available in standard pin packages as well as the newest Synergy patented metal surface mount package with either leaded or flush mounting configurations.

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INFO/CARD 24

Mixed Bag for Oscillator and Crystal Bookings — The Electronic Industries Association Piezoelectric Devices Division has reported that quartz oscillator bookings for December 1992 were up 3.9 percent in dollar value but down 5.8 percent on the number of units booked compared to November. Year-to-date bookings were down 7.0 percent in dollar value but up 16.1 percent in units.

Crystal bookings for December were up 33 percent in dollars and 25.9 percent in units compared to November bookings. Year-to-date crystal bookings were up 6. 3 percent in units but down 5.8 percent in dollar value.

Sciteq Electronics Relocates — Sciteq Electronics, Inc. has announced its move to new facilities. Their new address is 4775 Viewridge Avenue, San Diego, CA 92123-1641. Their telephone and facsimile numbers remain the same.

Electro-Mechanics Company Sold — ESCO Electronics Corporation has announced the purchase of Hart Properties and its single operating entity, the Electro-Mechanics Company. ESCO paid $1.6 million for 100 percent of HPI's and EMCO's capital stock at closing and will assume a debt of approximately $3 million.

Sprague-Goodman Acquires Trimcon — Sprague-Goodman has announced the acquisition of Trimcon, a manufacturer of multiturn air dielectric trimmer capacitors and microwave tuners. The Sprague-Goodman affiliate will be known as SGE Microwave, Inc. Terms of the sale were not disclosed.

Subsidiaries Pool Talents — Two manufacturers of RF communications systems and components, The Antenna Specialists Co. and Decibel Products, both wholly owned subsidiaries of Allen Group, Inc., have become working partners within a new corporate structure to be known as the Allen Telecom Group. The two companies will continue to independently market their distinctive products under their traditional brand names but will otherwise pool their engineering, manufacturing and distribution strengths.

Micron Communications Formed — Micron Technology, Inc. recently announced the formation of a new subsidiary, Micron Communications, Inc. The new company will develop new products for the RF identification marketplace. Their address is 2805 East Columbia Road, Boise, Idaho 83706-9698. Tel: (208) 368-4000, fax: (208) 368- 4558.

LNR Awarded X-Band Contract — LNR Communications, Inc., has been awarded a contract to implement a turnkey, X-band communications link operating via the 7/8 GHz NATO/DSCS satellites and utilizing commercial off-the-shelf equipment. LNR will deliver, install, test and support a Hub Earth Station utilizing a nine meter diameter antenna as well as 2.4 meter Mobile Terminals. The terminals also comprise beacon tracking, RF/IF electronics, digital modems, uplink power control, and a centralized monitor and control system. Financial terms of the contract were not released.

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INFO/CARD 26

Cellular Systems — Reaching Out With RF

By Gary A. Breed
Editor

At this instant in history, cellular radio-telephone systems are among the top few markets for RF products. And, with implementation of more advanced dual-mode and digital cellular systems, it will continue to be extremely important to the RF industry.

The recent RF Expo West was a good example of cellular's importance. Technical papers on the subject addressed designs for GSM (the Pan-European digital cellular system), microcell system considerations, new power amplifier technology, and speech encoding methods. If you include other wireless telecommunications systems that use cellular technology as a springboard, the sum includes over half of the conference papers.

Further evidence of the importance of cellular is the litany of new products introduced for this application. A quick summary of recent new products includes miniature SAW filters from Hitachi, TCXOs from TEW, base station power transistors from SGS-Thomson, IF integrated circuits from Signetics, synthesizers from National Semiconductor, amplifier modules from Q-Bit, ceramic resonators and filter elements from Trans-Tech, hybrid transformers from Sage Laboratories and a dual-mode chip set from AT&T Microelectronics. Of course, this list mentions only a few of the most recent announcements. Any comprehensive list of RF components for cellular applications would fill several catalogs!

The forecast for cellular is summed up by Gary LaBelle of Hewlett-Packard's Communications Components Division, "Analog cellular is a large market and holding its own." But, he adds, "We'll be seeing a step up in growth" as digital cellular systems enter full production. After saying this much, nearly everyone in the RF business expresses optimism that new technologies are in the works that will continue the strength that the cellular market has given to RF.

Miniaturization is a major effort in cellular equipment, such as these coaxial line elements from Trans-Tech.

Keeping engineers busy are the theoretical and practical challenges in manufacturing a cost-competitive product, and making it conform to increasingly complex operating protocols and performance demands that digital and dual-mode cellular systems require. GSM, NADC (North American Digital Cellular), JDC (Japanese Digital Cellular), CDMA and TDMA are only some of the technical variations on the transmission methods used in digital cellular. I and Q modulation and demodulation are key engineering concepts, along with filter and amplifier linearity, digitally-encoded voice and data information and digital signal processing.

Manufacturing is a Major Factor

As a mature technology (analog cellular) and a generally well-defined system (digital cellular), the majority of RF engineering time is being spent on manufacturing-related problems. Size and power consumption are clearly the primary forces driving this part of cellular product engineering. In fact, cellular products may be getting the greatest design-for-manufacturing effort in all of consumer electronics, if we consider a recent study done at Storage Tek.

Kathy Baker headed a team of Storage Tek engineers who evaluated a wide range of consumer electronics devices — notebook computers, palm-size digital assistants, portable CD players, video cam-corders and cellular phones. Each product was disassembled and analyzed to compare different companies' techniques and to identify trends in consumer manufacturing that could lead to better manufacturing processes at Storage Tek.

One of the conclusions reached by Baker's group was that a cellular phone, specifically, Motorola's Micro-TAC unit, used the most advanced techniques among the products examined. The Micro-TAC uses a six-layer printed wiring board with 5-mil trace width, hundreds of blind vias (layer-to-layer plated-through-hole conductors), together with an assembly technique that mounts components to both sides and solders them in a single pass. A further reduction in manufacturing operations is made by innovative use of the enclosure for mechanical support, holding locating pins for alignment, applying pressure for elastomeric connectors, as well as being conductive for EMC performance. Finally, a single screw holds it all together, and it is hidden beneath a label on the finished product.

The attention to detail and the careful way many functions and operations are combined in the mechanical assembly also applies to the electronic portion of cellular products. The reason we see a continuous stream of new products is because the demands for small size and increased functionality must be achieved simultaneously. It is probably fair to state that the concept of a "mature" market is no longer valid. Even in an established market like cellular telephone equipment, there is a constant effort to respond to new demands and to achieve greater performance and convenience. **RF**

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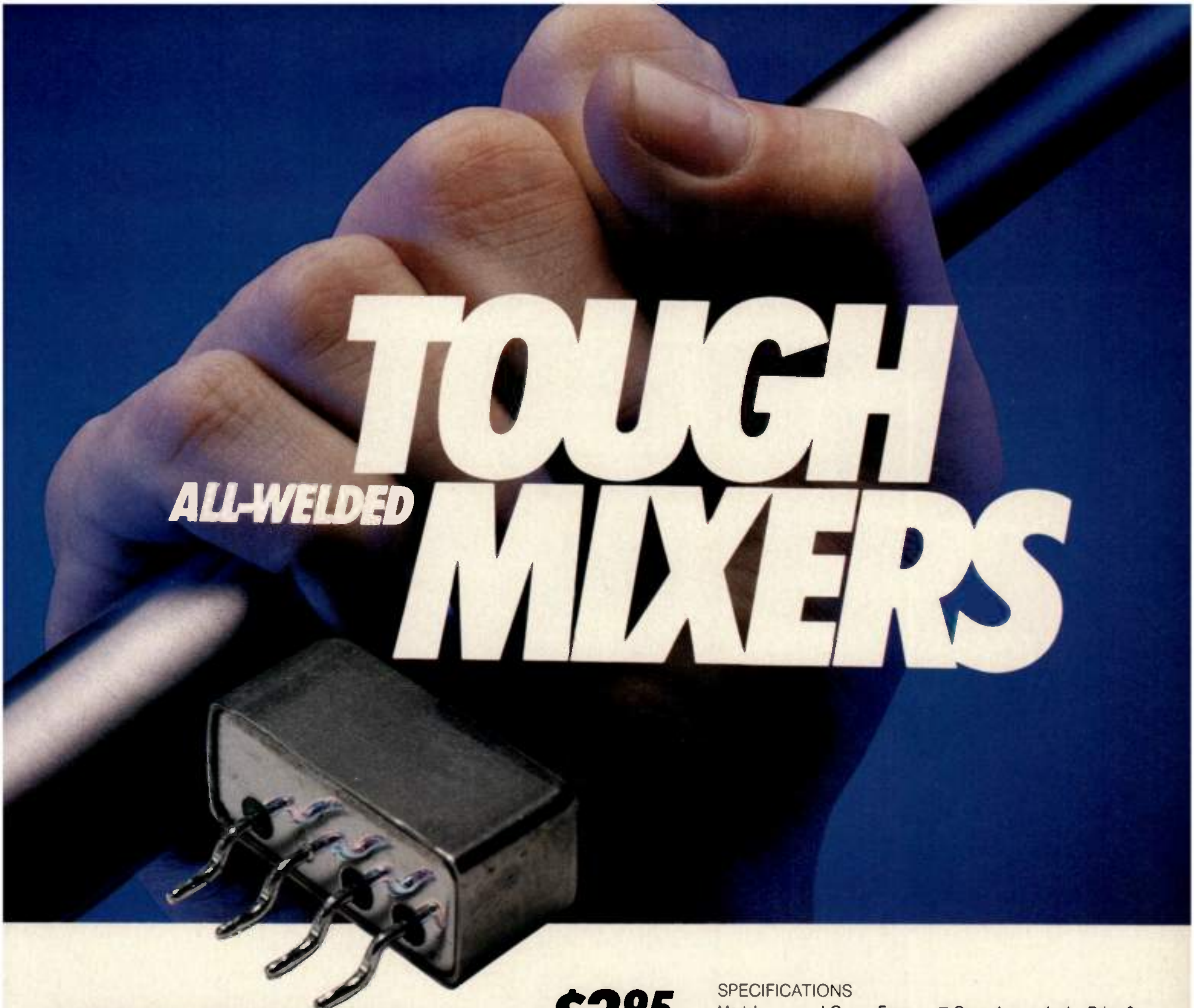

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SPECIFICATIONS

Model	LO Power (dBm)	Freq. LO/RF (MHz)	■ Conv. Loss (dB) $\overline{X}$	δ	Isol. L-R (dB)	Price,$ Ea. 10 qty
TUF-3	7	0.15-400	4.98	0.34	46	5.95
TUF-3LH	10		4.8	0.37	51	7.95
TUF-3MH	13		5.0	0.33	46	8.95
TUF-3H	17		5.0	0.33	50	10.95
TUF-1	7	2-600	5.82	0.19	42	3.95
TUF-1LH	10		6.0	0.17	50	5.95
TUF-1MH	13		6.3	0.12	50	6.95
TUF-1H	17		5.9	0.18	50	8.95
TUF-2	7	50-1000	5.73	0.30	47	4.95
TUF-2LH	10		5.2	0.3	44	6.95
TUF-2MH	13		6.0	0.25	47	7.95
TUF-2H	17		6.2	0.22	47.	9..95
TUF-5	7	20-1500	6.58	0.40	42	8.95
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TUF-11AH	17		7.3	0.28	35	19.95

*To specify surface-mount models, add SM after P/N shown.
■ $\overline{X}$ = Average conversion loss at upper end of midband ($f_U/2$)
δ = Sigma or standard deviation

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Simple, Low-Cost UHF VTOs

By Andrzej B. Przedpełski
The SHEDD Group

There are times when, either because of cost considerations or standard component nonavailability, you have to design your own VTO for a specific application. A simple voltage tuned oscillator (VTO) using a MMIC and a transmission line for the feedback circuit may be the answer for non-critical applications. This simple oscillator design can be used in a narrow range PLL.

Most of the standard circuits using discrete transistors are quite complicated and hard to implement. But, there is an easier way! Using standard low-cost 50 ohm amplifier devices and techniques similar to those used in the crystal oscillator designs (1), VTOs for specific tuning ranges can be easily designed. "Designed" may be the wrong word to describe this process, since it is easier to optimize the circuit by trial-and-error than by rigorous analysis. However, because of the simplicity and noncritical nature of the basic circuit, this process is fast and the success ratio is high.

Starting with the Barkhausen oscillator requirements — i.e. the open loop gain of the feedback circuit has to be more than unity, and the phase shift has to be an integer (including zero) multiple of 360 degrees — it is easy to visualize the required circuit. This is shown in Figure 1. The two requirements are then:

$$G(s) \times H(s) \geq 1 \text{ at an angle of } 360n \text{ degrees} \quad (1)$$

where n = 0,1,2,3 etc.

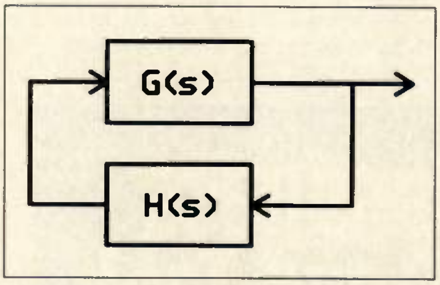


Figure 1. Feedback oscillator model.

The simplest implementation is the use of a 50 ohm MMIC or hybrid amplifier for G(s) and a piece of transmission line for H(s). The amplifier will provide the required gain while the transmission line provides the phase shift necessary to give the required total open loop phase shift.

The commercially available modular amplifiers are essentially either inverting (180 degree phase shift) or non-inverting (zero degrees phase shift) types. However, at the frequencies of interest, their phase shifts are typically 150 and –30 degrees respectively. For the designs described here, the inverting type was used. The main reason for this decision was better availability of the inverting type.

First, a suitable amplifier was selected. For these tests the Avantek GPD-401 was used for three reasons: it is a low cost device, it has built-in input and output capacitors, and several were left over in the lab from previous projects. The device uses hybrid construction, which allowed the manufacturer to build-in internal coupling capacitors. If the usual MMIC device is used, these capacitors have to be provided externally, where necessary, for DC isolation.

Looking at the S21 parameter on the data sheet, it is evident that this is the inverting type. In the frequency range of interest, the gain is about 15 dB and the phase shift is 160-120 degrees (in a 50 ohm system). Unfortunately, these figures only apply to class A operation. By its nature, the oscillator operates in a partially saturated condition, where the gain is lower. This provides the self limiting action. Both S21 and S11 are time-varying and are not defined. Avantek engineers suggest (2) using the SPICE models given for the MMIC devices. They feel that it is almost impossible to predict how S21 and S11 will change, since their change, under overdrive conditions, is a function of internal configuration. The only known change is the reduction of the magnitude of S21 by the amount of compression (the S21 angle change should be small for moderate compression levels). However, the general feeling is that the phase will not change appreciably for small amounts of overdrive.

Fortunately, it is easy to compensate for all these effects by the transmission line used for the feedback. The length of this transmission line can be varied experimentally until the required frequency is obtained.

The actual circuit used for the breadboard tests is shown in Figure 2. Its simplicity is quite apparent. Capacitors C1 or C2 are the tuning elements, and standard varactor diodes can be used to provide the needed frequency coverage. The transmission line, T, is the main feedback

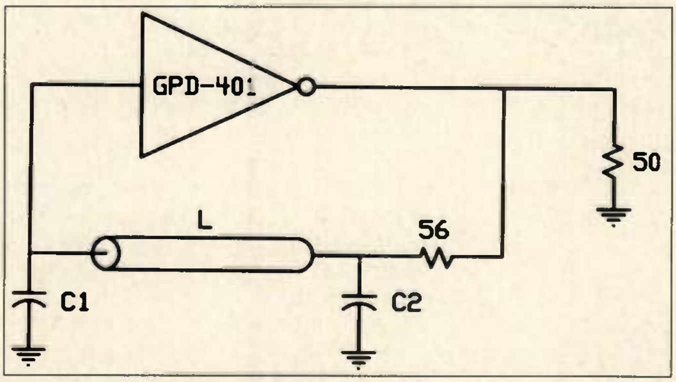


Figure 2. Oscillator circuit.

T Length	C1	C2	FREQ	OUTPUT
5″	0	—	410 MHz	+2 dBm
5″	3 pF	—	370 MHz	+2 dBm
1.9″	0	—	590 MHz	+5 dBm
1.9″	3 pF	—	530 MHz	+5 dBm
0.9″	0	0	660 MHz	+6 dBm
0.9″	3 pF	0	620 MHz	+6 dBm
0.9″	0	3 pF	570 MHz	+4 dBm

Table 1. Results of experimentation on circuit values.

phase shifting element. The resistor, R, has two main functions: it provides some isolation of the output from the feedback circuit and adjusts the loop gain to the needed value. The open loop gain should be more than unity to provide reliable start of oscillations. However, unnecessarily high gain increases the level of harmonics. This may be desirable in some cases, but in general, lower values of gain are more desirable.

To reduce the pulling effect of a changing load, a second amplifier stage can be used to provide load isolation and improve frequency stability.

As mentioned before, the easiest method of "designing" the desired VTO is to build it experimentally. The process is very simple and does not take much time. Table 1 shows experimental results using the circuit of Figure 2. Different lengths of standard 50 ohm coax were used and either C1 or C2 was "tuned" to get a feel of performance. After the desired range is obtained, the circuit can be constructed on a PC board using a 50 ohm microstrip transmission line for the feedback path. Thus, a low cost VTO, suitable for use in a narrow tuning range PLL, can be constructed.

A 50 ohm transmission line was used because of the ready availability of 50 ohm coax. Other impedances can be used, especially if microstrip construction is employed. The feedback loop gain and phase shift, H(s), will change, however.

While probably the easiest method of determining and optimizing the circuit constants is to construct a breadboard, the circuit can also be theoretically analyzed by using component s-parameters. The open loop gain, G(s) × H(s), can be determined by calculating the s-parameters of the cascaded individual components (4) and then using the standard gain formula (5). Some feel for performance can also be obtained through the use of a network analysis program (3).

The active device s-parameters can be obtained from the component data sheet. These will be for class A operation and may not be representative. However, if better data is not available, this may be a good starting point.

The transmission line s-parameters can be calculated for any line impedance.

$$S11 = S22 = \frac{j\tan\phi\ (Z/50 - 50/Z)}{2 + j\tan\phi\ (Z/50 + 50/Z)} \quad (2)$$

$$S12 = S21 = \frac{100Z}{100Z\cos\phi + j\sin\phi\ (Z^2 + 50^2)} \quad (3)$$

Where ϕ = electrical line length in degrees.

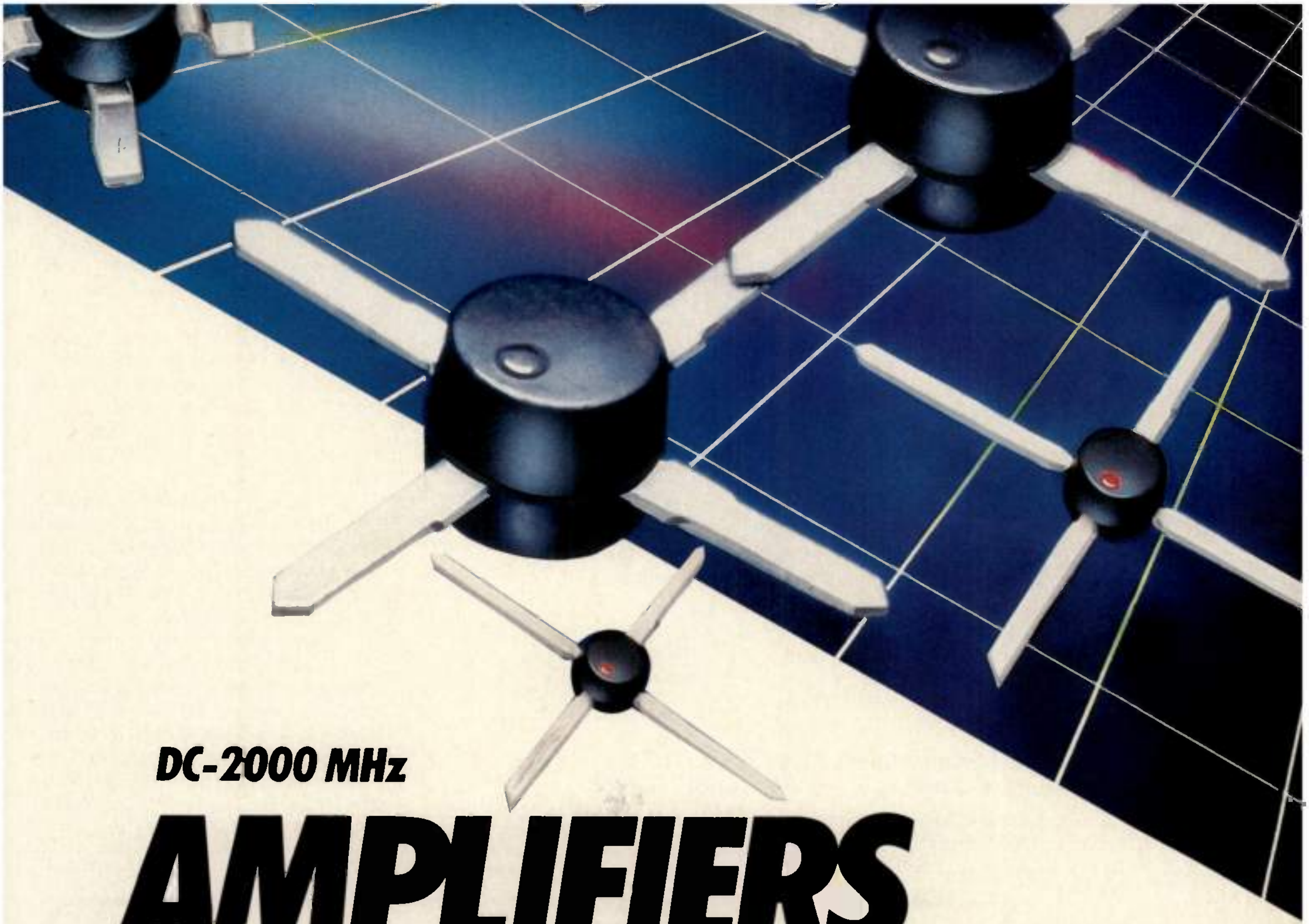
For a 50 ohm transmission line these simplify to:

$$S11 = S22 = 0 \quad (4)$$

$$S12 = S21 = 1 \text{ at an angle of } -0 \text{ degrees} \quad (5)$$

The main problem with trying to calculate the circuit performance is the stray phase shifts caused by finite lead lengths and stray reactances. This is especially true at the higher frequencies.

It should be kept in mind that this is a low Q circuit. Thus, it is not a very stable circuit by itself. However, for fast switching PLL applications it may be



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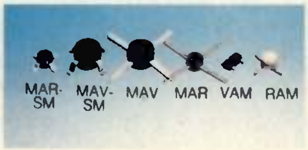

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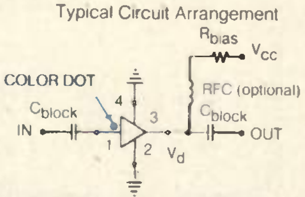
Typical Circuit Arrangement



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quite satisfactory; and it is simple and low cost! No spurious outputs or excessive noise were observed. **RF**

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About the Author

Andrzej (Andy) Przedpelski is the consulting editor for *RF Design* and founder of the SHEDD Group, an RF design, military product consulting group. He can be reached at 7260 Terrace Place, Boulder, CO 80303-4638, or by phone at (303) 499-9517.

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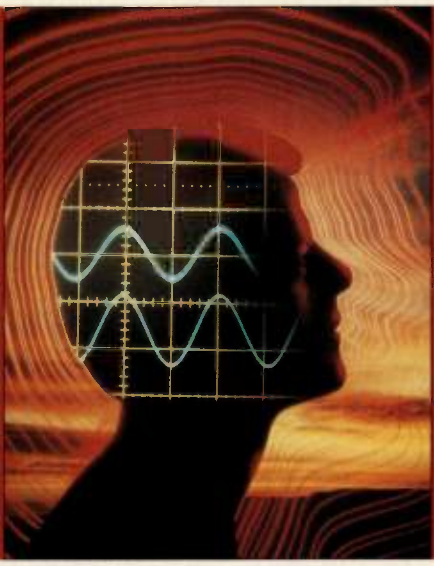

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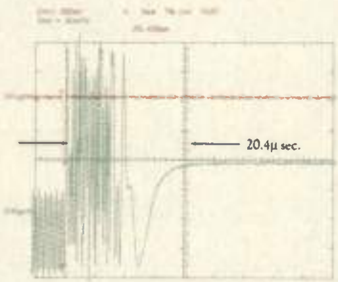

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Eliminating Spurious Responses in Phase Noise Measurements

By Bruce Long, P.E.
McCoy Electronics

Phase noise plots always seem to have a few spurious signals at integer multiples of the line frequency. These line frequency related responses are usually attributed to instrumentation ground loops and power supply hum and are quietly ignored. However, convincing the customer or quality control inspector they should ignore them too may not be so easy.

The central difficulty in phase noise measurement is accurate measurement of a small amount of noise against the backdrop of a much stronger carrier signal. Noise sidebands of a quartz controlled source measured in a one hertz bandwidth can be 170 dB weaker than the carrier. It's difficult to visualize ratios as large as ten to the seventeenth. Let's start with a large commonplace quantity; the total miles driven by Americans on a typical day. According to the Department of Energy, national gasoline consumption for 1991 was about 300 million gallons per day. A fleet average mileage of 25 miles per gallon gives 7.5 billion daily miles. One part in ten to the seventeenth gives a five thousandths of an inch. Phase noise measurement can be compared to monitoring the movement of a single automobile through a distance equal to the thickness of a sheet of paper against all the miles travelled by all the automobiles in the United States in a day.

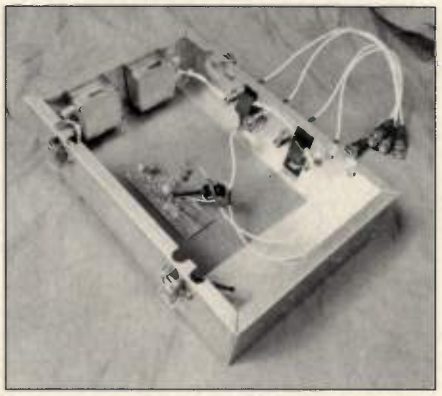

Figure 1. Phase noise test set-up.

Given the weak signals involved, it's not hard to understand why phase noise testing is highly susceptible to external interference. Coherent interference shows up as a line or spike on the phase noise plot. Strong spurious signals activate the spectrum analyzer auto-ranging function, sometimes falsely raising the noise floor. Multiple, weaker spurious signals merge together masking legitimate plot details.

The most troublesome spurious responses seem to be line frequency related. Believing ground loops and power supply hum to be the cause, I tried ground straps, isolation transformers, and large electrolytic capacitors in an unsuccessful effort to eradicate line related signals. I did notice changing the length of the RF coaxial cables between the units under test and the input ports of the phase noise instrument sometimes changed the line spurious levels. A few extra inches of transmission line might affect RF performance but it shouldn't have much effect at sixty Hertz. Could some line frequency spurs be a radio frequency phenomena?

Amateur radio experience sheds light on this issue. Direct conversion receivers are a favorite home construction project. Direct conversion receivers heterodyne RF signals directly to baseband without using an intermediate conversion which gives respectable performance without a lot of circuitry. Direct conversion receivers have one dramatic shortcoming — having high gain at audio frequencies, they are extremely susceptible to hum. On the surface this appears to be a simple issue of power supply ripple rejection but even battery powered direct conversion receivers suffer from hum and the hum often gets worse upon connecting antenna and earth ground.

The hum generating mechanism is rather sneaky. Radio frequency energy from the local oscillator travels along the supply leads into the power supply. Some energy finds its way to the rectifiers. The rectifier diodes, driven in and out of conduction at the line frequency, chop the stray RF, creating modulation

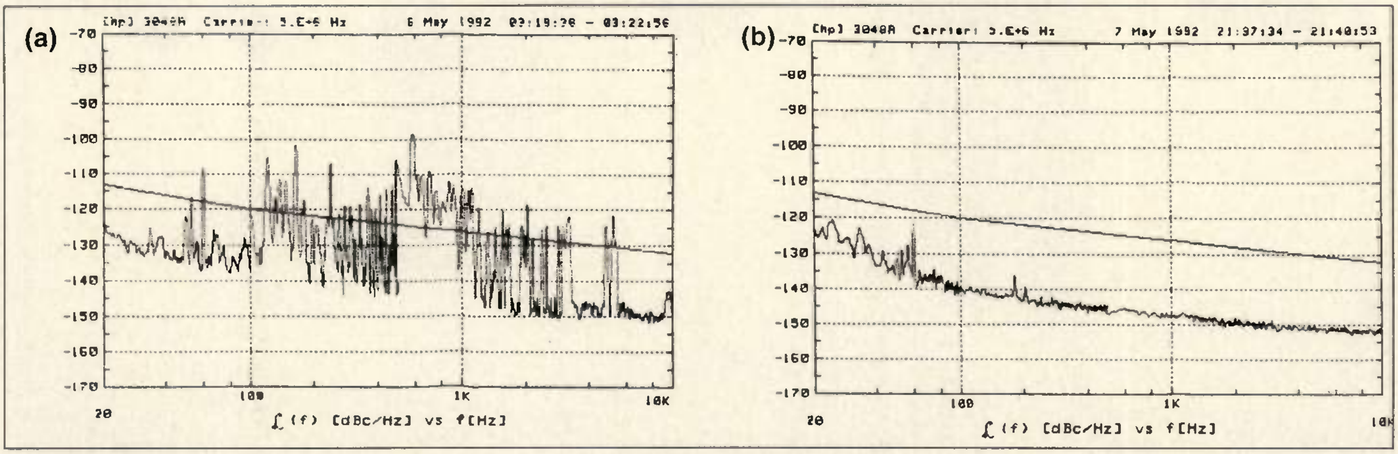

Figure 2. Phase noise plot of a crystal oscillator a) without box, b) with box.

sidebands. Modulated RF leaks back into the receiver, mixes with the local oscillator, and the receiver hums loudly despite a ripple free power supply.

A battery power supply helps but may not be sufficient if the receiver has inadequate shielding. Radiation rather than conduction is the villain here. The local oscillator must be inside an RF tight enclosure with DC lines decoupled and bypassed. An RF amplifier with good reverse isolation and a balanced mixer reduces leakage via the antenna and sometimes it helps to lowpass filter the audio output to prevent leakage by that route. These preventive measures make hum-free direct conversion reception possible.

Phase noise test systems and direct conversion receivers are similar in several respects. Both convert sideband energy to baseband, both have a lot of audio frequency gain, and both are susceptible to RF induced line frequency hum. Shielding and decoupling are the key to reducing line related phase noise spurs. The goal is to keep RF energy out of line driven rectifiers. Switching power supplies and display monitors are also potential hum modulators causing spurious signals to appear

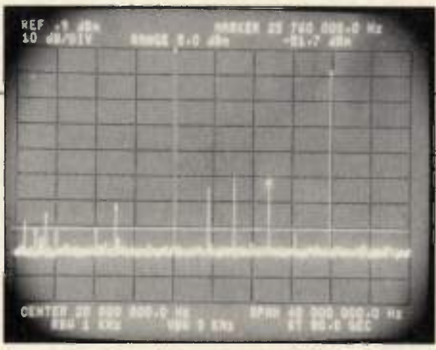

Figure 3a. Harmonic level measurements without box.

at the switching and scanning frequencies. Actually a multitude of hum modulators must be considered given a typical phase noise system has a monitor, a computer, at least one spectrum analyzer and the likelihood of other nearby unrelated test equipment.

Figure 1 shows a simple solution — a shielded and filtered metal box for phase noise measurements. The units under test go in the box. Power line filters, contained within separate shielded cans, decouple the power supply. Coaxial cables connect the shield box to the phase noise test set. Female-to-female bulkhead coaxial connectors can be used to pass RF and tuning voltage signals through the test box wall, but short

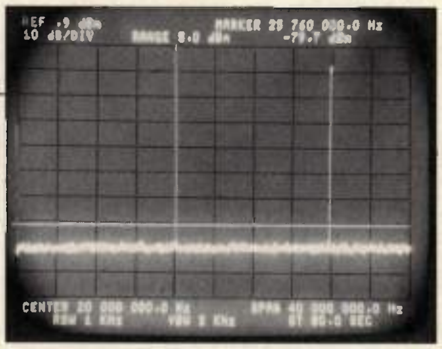

Figure 3b. Harmonic level measurements with box.

test cables always seem to be scarce so I used female bulkhead crimp connectors facing inward with short RG-188 BNC pigtails permanently attached.

Snap-action latches secure the lid. A previous version used captive threaded fasteners and thumb screws but tightening and removing six screws with every phase noise test quickly inspires other solutions. The latches are more convenient but take a little forethought in placement. Because they pull down at the very edge of the thin aluminum lid, they tend to bow the lid upwards in the center and in between the latches. A thicker lid, internal bracing, conductive gaskets, or a little less latch force come to mind as improvements for a third generation shield box, but in many cases the box works fine even without the lid.

Inside the box are clip leads connected via feedthrough filters to banana jacks on the box ends. I used coax for these power supply runs inside the box with the braid connected to the chassis at the feedthrough filters to keep RF pick up to a minimum. The feedthrough filter on the first shield box I built consisted of two feedthrough capacitors and a bifilar choke wound common mode style on a high permeability junkbox toroid. On the second shield box, I used commercial feedthrough filters. Both work well.

A closed metal box exhibits waveguide type resonances at frequencies having wave lengths approaching the box dimensions. These can encourage instability especially in un-shielded, "dead bug" style breadboards. Sheets of black anti-static conductive foam glued to the walls, floor, and lid breakup unwanted resonances.

Figure 2 shows the phase noise plot of a crystal oscillator particularly suspectable to line frequency related spurious signals. The improvement with the box in place is dramatic. The shield box is also helpful for controlling injection locking during phase noise testing although two shield boxes — one for each oscillator — might be necessary.

Other radio frequency measurements benefit from the shield box. Figure 3 shows before and after results of a harmonic level measurement. External interference, visible without the shield box, is completely eliminated. In this application the shielding integrity of the spectrum analyzer comes into question. Modern test equipment with internal microprocessors, video displays, and switching power supplies must have excellent internal shielding to escape self-interference, making the unit under test the most likely entry point for external interference. Placing it in a shielded, filtered box can work almost as well as doing the measurement inside a screen room, but the shield box is inexpensive and portable. ***RF***

Acknowledgements

I'd like to thank Chuck Skovira, Manufacturing Engineer, McCoy Electronics, for suggesting the snap latches and making the second generation shield box.

Parts and Reference

Snap Latch
Neilsen Hardware, 770 Wethersfield Ave., P.O. Box 568 Hartford CT 06141-0568. Tel: (203) 522-8145, Fax: (203) 525-0180.

Feedthrough EMI filter
Murata Erie part number 9200-300-0025
Atten. 44 dB @ 1MHz
60 dB @ 10 MHz
70 dB @ 1GHz

Common Mode Hum
Amateur Radio Handbook, 1985, Sixty-second Edition, American Radio Relay League, Newington CT, Page 12-8.

About the Author

Bruce Long is a Senior RF Design Engineer for McCoy Electronics Company. He also recently passed his Professional Engineering Registration Exam and is now PE 043157E.He can be reached at 100 Watts Street, PO Box B, Mount Holly Springs, PA17055. Tel: (717) 486-3411, Fax: (717) 486-5920.

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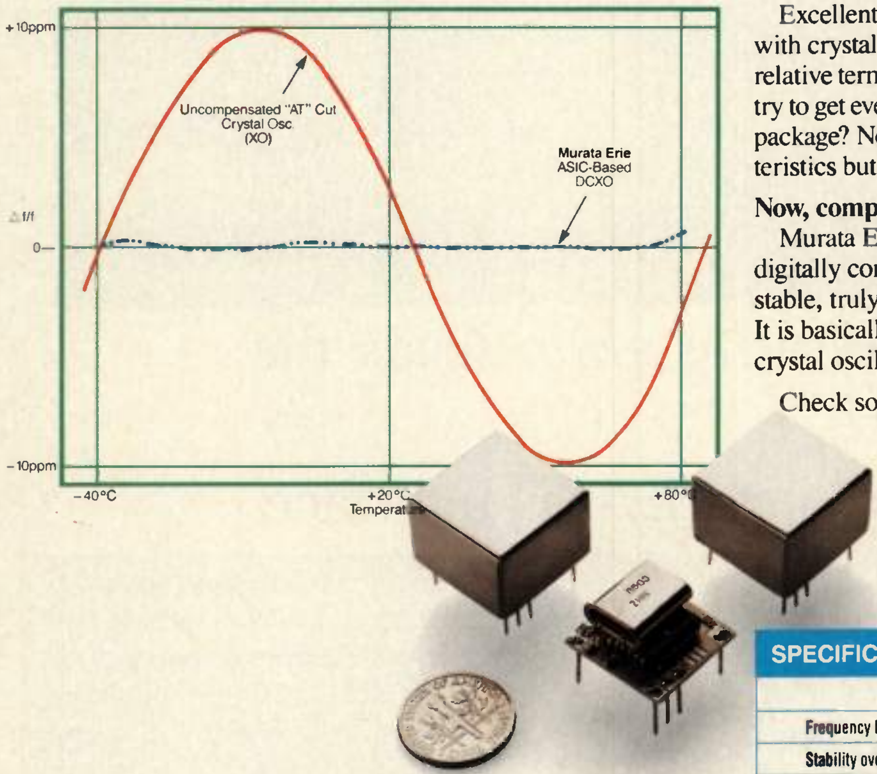

Excellent frequency stability has been a fact of life with crystal oscillators for years. So has small size (a relative term), low cost and flexibility. **But**, did you ever try to get everything that you needed in a single oscillator/ package? No way. One, maybe two desirable characteristics but certainly not all. You had to compromise.

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

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SPECIFICATIONS

	DC2210 AH
Frequency Range	10MHz to 25MHz
Stability over −40°C to +85°C	±0.5ppm
Output	"HC" CMOS
Power Dissipation	+5V @ 15mA
Short Term Stability	1×10E−9 @ T=1 sec.
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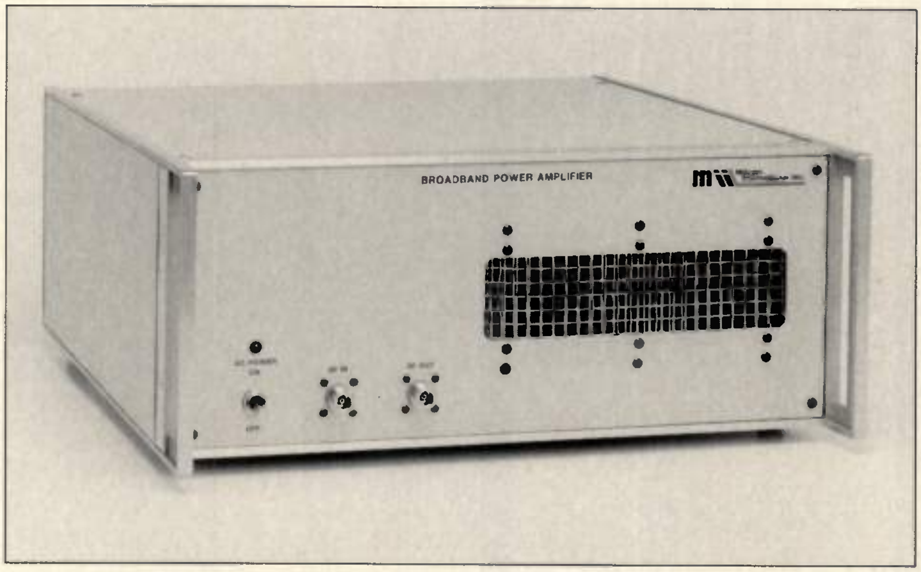
Class A Design Techniques Create a New Broadband Amplifier

By Charles Gentzler
Milcom International, Inc.



Since the introduction of the first internally matched RF power bipolar transistor in the early 1970's, the race to achieve more power over broader bandwidths has never ceased. The first devices were designed under military contract to replace vacuum tube transmitters that required mechanical tuning to cover the 225-400 MHz military communications band. These devices were designed for class C or lightly biased class AB operation. With the advent of improved device processing, the DC safe operating area improved to the level which permitted full class A operation. These high power devices spawned a new industry: The class A instrumentation amplifier. Amplifier companies saw the need for broadband component and system testing and introduced new class A linear amplifiers to meet the demands of a growing industry. The following article discusses some of the techniques used to generate broad band class A power.

The first solid state design technique was a spin-off of the growing CATV industry. These applications used a relatively small bipolar transistor with negative feedback (Figure 1) in a 1-5 watt building block, obtaining reasonably good input and output return loss and gain flatness. As many as 64 of these stages were combined for powers of over 100 watts at bandwidths in excess of 0.1 to 110 MHz (Figure 2). The design philosophy was solid and since modules were combined, it was easy to offer a product in many power levels by changing the number of combined modules. The basic power device had an F_t of several GHz, so there was enough loop gain to maintain good gain and phase match from module to module, a production manager's delight. The basic

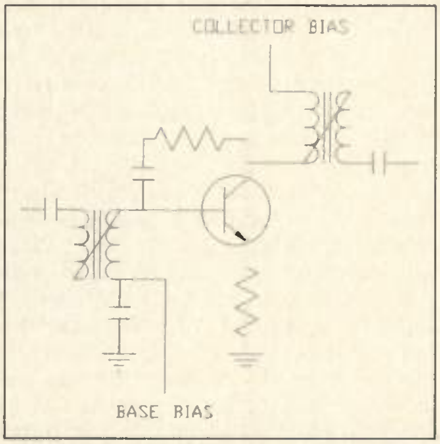


Figure 1. CATV style medium power amplifier with feedback.

TECHNICAL SPECIFICATIONS
MILCOM INTERNATIONAL MODEL LSA1051-50A
MOSFET DISTRIBUTED AMPLIFIER

FREQUENCY RANGE:	1 - 512 MHZ
POWER OUTPUT:	50 WATTS (MIN) CLASS A
LOAD VSWR, RATED OPERATION:	<1.5:1 AT 50 WATTS (Stable into any load)
INPUT VSWR:	<1.5:1 (TYP) <2:1 (MAX)
OUTPUT VSWR:	<2:1 (TYP) <3:1 (MAX)
POWER INPUT:	0 DBM (TYP)
GAIN:	47 DB (MIN)
GAIN FLATNESS:	1.0 DB (TYP) 1.5 DB (MAX) (All power levels to 50 watts)
HARMONICS:	–25 DB (MIN) @ 50 WATTS
SPURIOUS:	–80 DB (MIN)
AC INPUT POWER:	115 VAC, 50/60 HZ
OPERATING TEMPERATURE:	0 C. TO +50 C.
STORAGE TEMPERATURE:	–35 C. TO +70 C.
RF CONNECTORS (IN/OUT):	TYPE N FEMALE
WEIGHT:	40 LBS
PHYSICAL DIMENSIONS:	19" RACK, 7" HIGH, 20" DEEP
COOLING:	INTERNAL FORCED AIR

Table 1. Specifications for the new broadband amplifier.

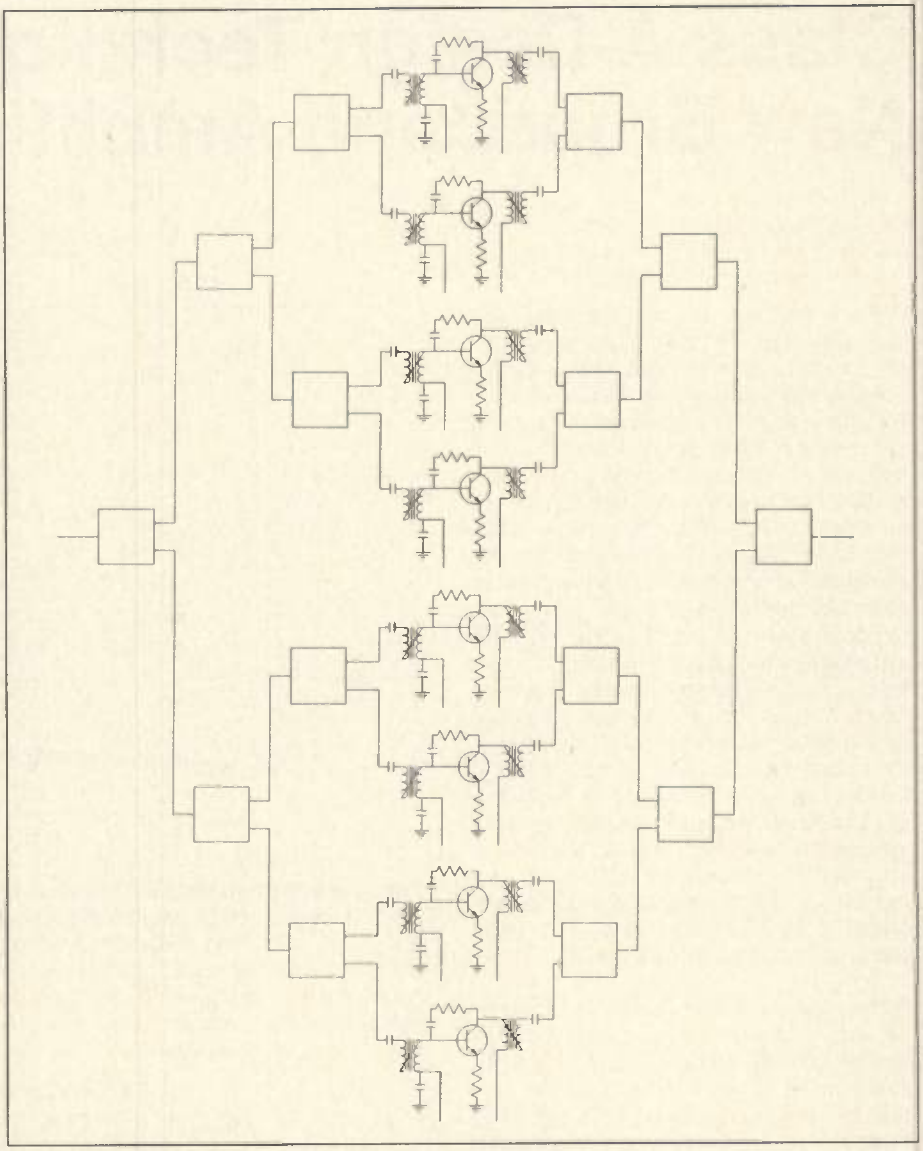


Figure 2. Multiple-stage combining technique.

limitation was the fact that even the best combiners have losses which increase with frequency. When trying to achieve broadband performance into the UHF band, the combining losses become significant. For example, a 1-500 MHz broadband combiner may have less than 0.1 dB loss at 1 MHz, while the loss at 500 MHz may be as much as 1 dB. With a cascade of several combined stages the available power at the higher frequencies will drop off significantly.

Another approach to broadband power design was to use a pair of bipolar devices to develop the power in one stage (Figure 3). First, one had to select a device with as low an output capacitance as possible and a large DC operating area for class A operation. The development of the heavily ballasted UHF 100 watt device made possible the development of good 25 watt class A stage operating up to 200 MHz. A pair of these devices were operated in push pull to achieve 50 watts class A linear power from 1-200 MHz. But circuit design and reproducibility was a challenge. Since the operating impedances are very low, gain sloping rather than negative feedback is the primary design element to achieve flat power gain versus frequency. The DC beta of the device primarily controls the low frequency gain and the F_t of the device controls the high frequency gain. Even with both parameters specified to some degree, individual test selection of gain

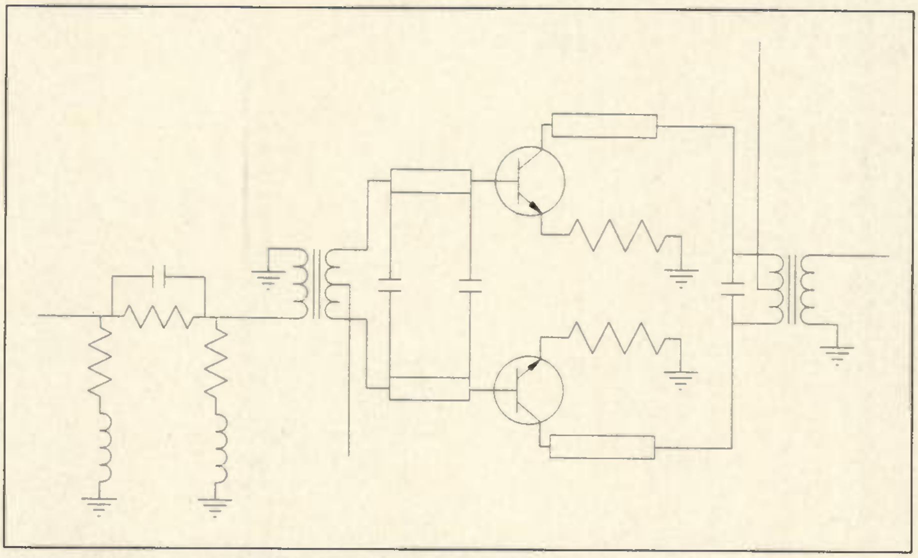


Figure 3. Push-pull bipolar class A amplifier.

sloping components was sometimes necessary, much to the dismay of the production manager. A primary advantage of this technique is that four 50 watt modules can be combined to easily obtain 200 watts output power without bandwidth shrinkage. Obtaining this performance with a 5 watt module would be extremely difficult. A major limitation of this high power approach is the ability of the amplifier to deliver full rated power into moderate VSWRs. The amplifier could deliver more current into a lower impedance, but the voltage swing of the load line limited the amount of power which could be supplied into a high impedance.

A major development in the evolution of broad band design was the introduction in 1977 of the first push pull package. The engineering community was quick to accept this new approach for improved broad band performance. A new market existed for an instrumentation amplifier that covered the newly introduced 100-500 MHz power devices. Using a similar design approach as outlined above, it was possible to design a 100-500 MHz class A stage with an output power of 25 watts. Since the band-

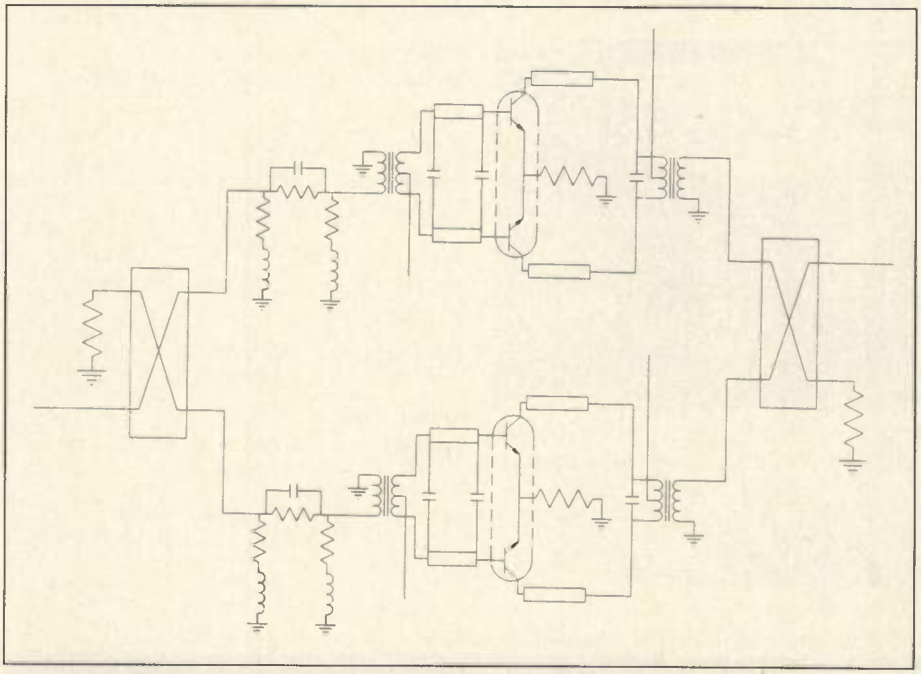


Figure 4. 50-watt building block using matched push-pull devices.

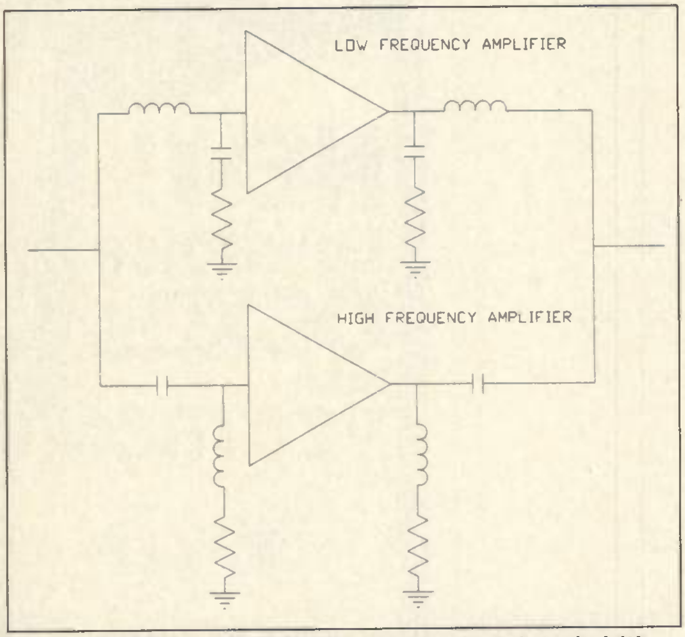


Figure 5. Diplexed amplifiers for greater bandwidth.

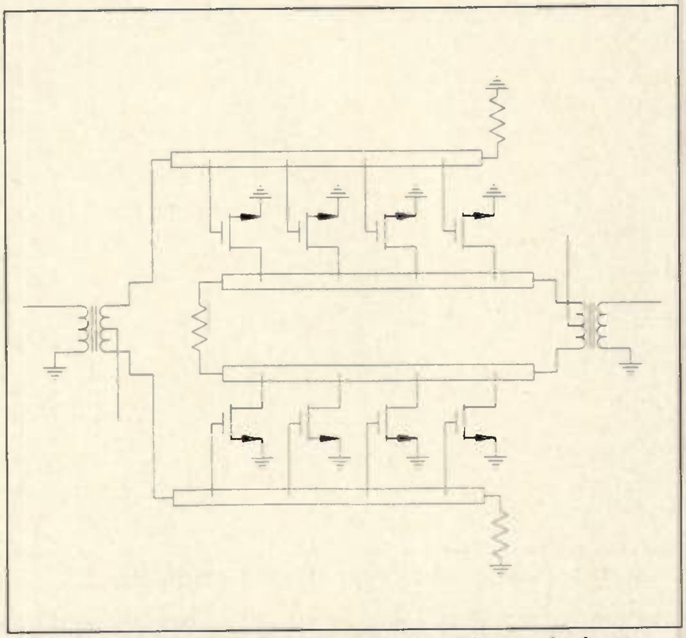


Figure 6. MOSFET distributed amplifier technique.

width of 5:1 can be accommodated with a three section quadrature combiner, it was possible to design a 40-50 watt building block with only two modules, as shown in Figure 4. The quadrature combiner, in addition to providing a combin-

ing function, provided a building block that had inherently low input and output return loss. Since the transistors were operated 90 degrees out of phase (with cancellation of equal return reflections), the module performed very well when driving moderate VSWRs. The resulting modules could then use conventional in-phase combining to easily obtain power in the 200 watt range. A similar technique was used several years later to achieve class A power in the 100 watt range from 500-1000 MHz.

The next major broadband milestone was the introduction of the high power FET. It basically had no safe operating area limitation and exhibited an input impedance of basically the same level as bipolar devices with input matching. This was a major development in the instrumentation marketplace because its volume is relative low compared to the communications and military market. The FET gave the designer the capability to optimize amplifier designs without begging for custom input matches which were generally required with bipolar designs. The elimination of secondary breakdown, coupled with the high DC input impedance, made class A biasing a design problem of the past.

Today, the most versatile design block is a dual push-pull design which is quadrature coupled. The push pull design exhibits improved broad band matching with even-order harmonic cancellation. Quadrature combining preserves the advantages of push pull design and adds input and output reflection cancellation, third harmonic reduction, and greatly improved load driving capability.

Unfortunately, a quadrature combiner requires a 90 degree phase shift in one of the input/output ports to function. This performance can be obtained in many structures. However, the object is to obtain a 90 degree phase shift over the broadest bandwidth as possible. A three section coupled symmetrical line structure is capable of 5:1 bandwidth at the expense of in-band ripple of ±0.75 dB. Lumped quadrature combiners have been designed with ripples as low as ±0.15 dB over the same bandwidth at the expense of some phase imbalance. There are even some designs in the industry which have approached 25:1 bandwidth at the expense of loss and lower power capability. These designs still are very useful because only the first building block needs quadrature combining. In-phase combining may be used with these building blocks to achieve high power levels while retaining all the benefits mentioned earlier.

To provide a preamplifier-driver using solid state technology for the broad band vacuum tube distributed amplifiers, a novel design approach was developed. Two amplifiers were diplexed to cover an extremely wide bandwidth (Figure 5). Since most solid state devices at power levels in the excess of 10 watts rely on RF transformer technology, the bandwidth of the transformers becomes a limiting factor, especially at the lower frequencies. To overcome this basic limitation, overlapping amplifiers were designed and coupled by diplexers to provide continuous bandwidths in the excess of 20,000:1.

A further market which has emerged can be called the "ultra-broadband" design. Amplifiers that cover from 1-500 MHz or even 1-1000 MHz, at power levels from a few watts to as high as 100 watts, find applications in fields as diverse as component testing and medical research.

Virtually all the designs start with a low power module, using combining technology to achieve quite respectable performance. The basic limitation is the loss slope of the combiner. As more modules are combined for higher power, the combiner loss at the higher frequencies creates a natural power roll-off with frequency. Typically, the maximum available power drops almost 3 dB from midband to the upper band edge due to combining losses.

Distributed Amplifiers

A broad band technology that has been around since the 1940s is the distributed amplifier. In the past, it has been almost the exclusive domain of the vacuum tube. Today there are a number of companies offering amplifiers supplying over 1000 watts from 10 KHz to over 200 MHz using this technology.

The basic principle of operation is to use the input and output capacitance of active devices as elements in a lumped transmission line. One end of the output line is a load to absorb reflections with the other end being the output. The input line is constructed in a similar

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fashion with the end of the input line also being terminated in a load resistor to absorb reflections. The properties of such an amplifier are broad bandwidth, good input (S11) and output VSWR (S22), high isolation (S12), improved load driving capability and relatively high power output. Attempts in the past to use bipolar devices in a distributed amplifier have not been very successful because of the low input impedance of power bipolar transistors. Emerging GaAs FET technology paved the way for solid state distributed amplifiers. There have been numerous circuits published and a number of commercial companies marketing such devices, either as a MIC or a complete amplifier. The power generally is less than 5 watts, but it is an accomplishment to develop several watts from 2-20 GHz. The same principles that make possible distributed amplifier operation with vacuum tubes and GaAs FETs can also be applied to silicon medium power FETs.

Milcom International has been successful in designing a single module which is capable of 50 watts linear operation from below 1 MHz to over 500 MHz (simplified diagram in Figure 6). This is accomplished by the use of ferrite technology to lower the operating impedance of the amplifier to a level where the input and output capacitances of the silicon FETs will function as the capacitive elements in a transmission line just as in the vacuum tube design. After the amplification process, the impedance is stepped back up to 50 ohms. To achieve higher power, the input signal is split in phase in addition to impedance transformation to allow push-pull operation. A module of this design retains the benefits of push pull operation, has improved capability to drive non 50 ohm load impedances over push-pull designs, and has excellent input and output VSWRs considering the amount of power being developed. Since the isolation of such a module is high enough that the load VSWR doesn't cause significant deterioration of the amplifier's input VSWR, a conventional push-pull preamplifier-driver can be employed. Another benefit is that 50 watts is being generated without traditional combining, and higher powers have been achieved with only a minimal high frequency loss, less than 0.5 dB. For general purpose instrumentation applications, this solid state high power distributed amplifier is an advancement in the state of the art of ultra bandwidth amplifiers.

Specifications of the Milcom LSA1051-50A broadband amplifier are shown in Table 1. Readers desiring more information can call Milcom International at (714) 554-1710, or circle Info/Card #250. **RF**

About the Author

Charles Gentzler has spent twenty years in the RF industry, designing broadband amplifiers and related products. He has been granted a patent in the area of biasing bipolar transistor for class A operation. He has held engineering management positions at Ailtech, EPSCO RF division, Lucas Zeta division, and currently is Vice President of Engineering at Milcom International, Inc., 10891 Capital Ave., Garden Grove, California 92643.

Miniature UHF Transmitter

Based on innovative Surface Acoustic Wave (SAW) technology, a new miniature UHF transmitter module ideal for a wide range of RF applications has been developed by RF Monolithics (RFM). Available in sample quantities by second quarter 1993, the HX1000 transmitter provides an excellent radio frequency alternative to many remote control applications that now use infrared (IR) technology. The first in the RFM HX series, the HX1000 operates at 433.92 MHZ. Its innovative package design and the proprietary SAW output filter implementation provide the capability to meet the stringent 1 nanowatt, 2nd harmonic emission requirement of the German FTZ 17TR2100 specification. The HX1000 is designed to operate from a 3 V lithium battery, and using a 10 percent duty cycle, will draw 0.75 mA average operating current. The transmitter measures 0.340 × 0.400 × 0.107 inches. The HX1000 is specified to operate over a temperature range of –40 to +85 degrees C with a frequency accuracy of 0.05 percent. It can support data rates of up to 3 kbps using pulse amplitude modulation. Production quantities will be available by fourth quarter 1993. Unit pricing is $7.63 in quantities of 10,000.

RF Monolithics, Inc.
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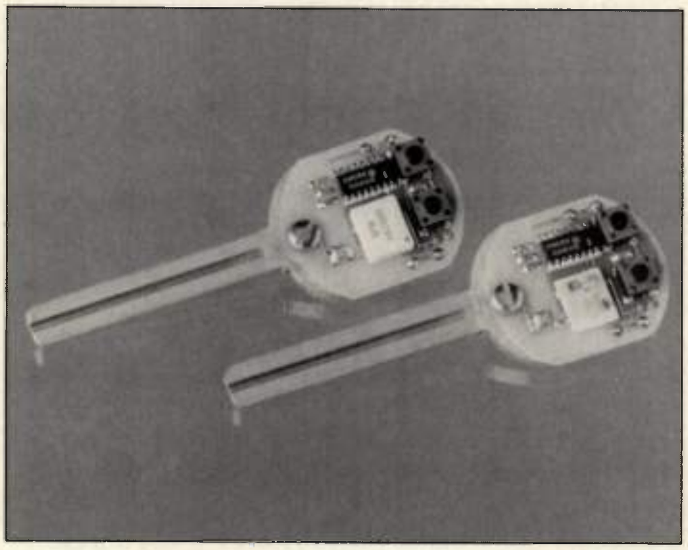


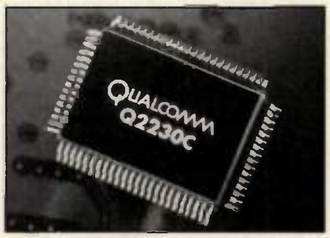
Low Frequency Crystals

TeleQuarz USA has announced a new family of low cost, low frequency crystals, the LFQ series. Built upon commercial "AT cut" technology, these crystals offer remarkably high performance at commodity level prices. The LFQ series crystal frequencies range from 2 to 14 MHz fundamental, with temperature stabilities to below 20 ppm at 0 to 70 degrees C. Aging rates of under 3 ppm/yr are typical. LFQ crystals are furnished primarily in standard HC-49/U, although other packages including shock resistant and SMD housings are available. LFQ crystals are suitable for both IR reflow and wave soldering techniques, and can be furnished in bulk or tape/reel format. A variety of standard frequencies are ready for immediate delivery; quantity prices start at under 20 cents per unit.

TeleQuarz USA
INFO/CARD #230

High-Speed Direct Digital Synthesizer

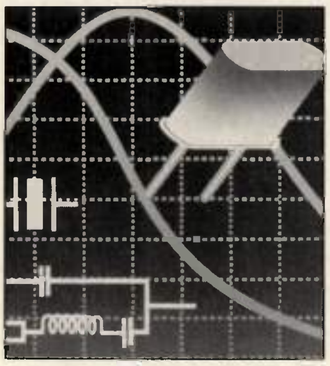
QUALCOMM, VLSI Products has announced the high-speed, single-chip Q2230 Direct Digital Synthesizer (DDS). The Q2230 DDS has a maximum system clock rate of 85 MHz, which allows it to synthesize output sinusoids from 0 up to 42 MHz when combined with a QUALCOMM Digital to Analog Converter (DAC). This synthesizer topology is perfect for applications that require fast frequency switching, good signal quality, and low cost. The Q2230 DDS consists of a frequency control interface, a 32-bit phase accumulator, a 15-bit sine lookup converter, and a 12-bit DAC output. Available in an 80-lead plastic quad flatpack package, QUALCOMM's Q2230 has a commercial operating temperature range of 0 to 70 degrees C. A single 5 V supply is required and the typical power dissipation is 980 mW operating at 85 MHz. Pricing for the Q2230C-85S1 is as low as $28.50 (quantity 1000). Availability is from stock.



QUALCOMM, Inc., VLSI Products Div.
INFO/CARD #229

Crystal Test Setup

An enhanced version of the SNA-120 crystal test setup from Wandel & Goltermann is now available, the SNA-120A. The new test setup uses a test technique which was recently incorporated into a IEC standard to be published as IEC 444-5. SNA-120A features include excellent accuracy and reproducibility of results, along with an enhanced test technique for measuring crystal equivalent electrical parameters in the frequency range above 125 MHz. A new test fixture allows the setup to measure LF crystals with excellent speed and accuracy. The advantages of the new fixture are most apparent when measuring LF quartz resonators with high series resistance. The test time is no longer a function of the settling time of the test object, making it possible to measure equivalent electrical data of clock crystals in less than 20 seconds.



Wandel & Goltermann, Inc.
INFO/CARD #228

Hand-Held Power Meter

The 6970 RF power meter, launched by Marconi Instruments, is the first RF and microwave power meter to provide accurate measurement capability in a rugged, hand-held portable unit. This cordless instrument covers 30 kHz to 40 GHz and provides similar functionality to currently available bench-top power meters at approximately half the cost. The 6970 uses the same power sensors as the Marconi 6900 series power meters and 6200 series microwave test sets. These sensors can measure power levels from –70 dBm to +35 dBm. Key features of the 6970 include a large three or four digit LCD display, an analog peaking indicator, and an audible pass/fail alarm to indicate when pre-set limits are exceeded. The 6970 weighs less than 1.25 lbs, and measures 3.5 × 7.5 × 1.75 inches. The 6970's built-in battery provides up to eight hours continuous operation and can operate from AC line while recharging. Price for the 6970 is $1522, with delivery in six weeks.

Marconi Instruments, Inc.
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RF DESIGN AWARDS CONTEST

Who Will Be the Winners?

Which software entry will win the Version 3.5 RF design software package for Windows from **EEsof, Inc.**?

Which design idea will win the UFX-BER bit-error-rate test instrument from **Noise Com**?

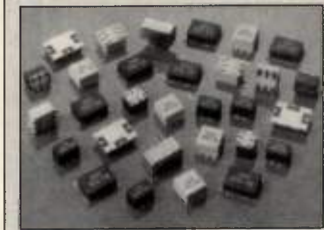
What great ideas will be published throughout the next year?

Watch for the results in the July issue of RF Design!

Product Spotlight: Mixers

Low Drive Mixers

The Starved L.O. series of mixers from Synergy Microwave features high quality performance with local oscillator drive levels of only +1 dBm. Series within the Starved L.O. line provide different packaging styles. All mixer series cover 0.5 - 2500 MHz, have typical conversion loss of 7 dB and typical RF/LO isolation of 35 dB. Prices for the standard surface mount package series, the SRD series, start at $6.30.



Synergy Microwave Corp.
INFO/CARD #226

SMT Active Mixer

The IAM-82008 from Hewlett-Packard covers the frequency ranges used for several services in the 900 MHz and 2.4 GHz bands. The active mixer offers conversion gain to 5 GHz and IF to 2 GHz. The IAM-82008 requires only external blocking capacitors for operation and can provide +8 dBm IF output, up to 15 dB conversion gain and requires only a 0 dBm LO signal. The device sells for $3.55 per piece in the thousands.

Hewlett-Packard Co.
INFO/CARD #225

High Dynamic Range

Locus produces high dynamic range mixer assemblies which cover the HF/VHF band with third order intercept points to +45 dBm. Image reject mixer downconverter and single sideband upconverter configurations provide 45 dB of image/sideband rejection. Wideband quadrature capability affords IRM/SSB basebands over a 20:1 bandwidth ratio. Supplemental IF filter configurations give over 90 dB of rejection.

Locus, Inc.
INFO/CARD #224

Low Intermodulation

The DMH-4R-1000 double balanced mixer from Merrimac uses two parallel ring modulators to produce a wide bandwidth circuit with very low intermodulation. Although designed for drive levels of +13 dBm, it will operate at drive levels from +7 to +17 dBm.

Merrimac Industries, Inc.
INFO/CARD #223

Two-Tone Performance

The model SBF0031M6/I3 MESFET mixer is designed for use as a down or upconverter in selected dense signal environments where dynamic range is critical. The high input intercept level of this mixer prevents unwanted cross modulation. Unlike Schottky diode mixers, only +26 dBm LO power is required for an unprecedented +36 dBm two-tone third order intercept point and +27 dBm (0.5 W) RF input 1 dB compression.

Miteq
INFO/CARD #222

SIGNAL SOURCES

SMT Clock Oscillators

AVX Corporation has introduced a line of AVX/Kyocera, surface mount clock oscillators. The AMO-HC series will drive up to 50 pF CMOS loads or 10 TTL gates. A de-coupling capacitor is built in, and the oscillators are available in several standard footprints and 15

standard frequencies.
AVX Corporation
INFO/CARD #221

Cellular/PCS Synthesizers

Synthesizers from RF Prototype Systems provide a simple low-cost solution for local oscillators in all the standard cellular/PCS and EAMPS cellular radios that use a common LO for both transmit and receive sections or those that use separate PLL's for transmit and receive sections. The synthesizers offer excellent phase noise, low noise and small size. Software for programming serial models is available parallel models are programmable with on-board DIP switches.
RF Prototype Systems
INFO/CARD #220

SMD, High Frequency

SaRonix announces a line of high frequency surface mount crystal controlled oscillators.

Available from 48 to 130 MHz with ACMOS output capable of driving both CMOS and TTL loads, the line boasts low input current and is available in several surface mount packages. The tri-state output is short circuit protected.
SaRonix
INFO/CARD #219

±0.25% Deviation VCXO

Vectron has developed a series of VCXOs and TC.VCXOs to provide deviation capability of ±0.25 percent while maintaining linearity to ±1 percent. The CO-271/275, CO-351/352 series are available in frequencies from 3.5 to 600 MHz. Price for a 20 MHz VCXO with ±0.25 percent deviation and ±0.01 percent 0/50° C temperature stability is $211 for 100 pieces and 10 week delivery.
Vectron Laboratories, Inc.
INFO/CARD #218

SIGNAL PROCESSING COMPONENTS

Failsafe Switches

Alan Industries announces two new series of high frequency coaxial failsafe switches. They are the SS series for the DC to 18 GHz range and the HS series for the DC to 12.4 GHz range. After an interruption of control voltage these failsafe switches return automatically to their original position. The switches are SPDT, and switching speed is 20 ms at 20 degrees C. Prices start at $215.
Alan Industries, Inc.
INFO/CARD #217

Bandpass Filter

KeLcom bandpass filter model BP-19.2/X16 offers phase linearity of < ±1 degree from 16.7 to 21.7 MHz. A minimum 3 dB bandwidth of 16 MHz centered around 19.2 MHz yields < 0.5 dB insertion loss with VSWR of 1.5:1. Outline dimensions are 0.8 × 0.4 × 0.4 inches with PC mount.
KeLcom
INFO/CARD #216

4-way SMT Divider

The Merrimac PDG-4E series of 4-way in-phase power dividers/combiners cover a broad frequency range of 2 to 2000 MHz. Versions are available in various broad bandwidths including 2-100, 10-500 and 100-2000 MHz. The compact package measures 0.8 × 0.8 × 0.17 inches and is hermetically welded for high reliability.
Merrimac Industries, Inc.
INFO/CARD #215

Plug-In Filters

Lorch Electronics offers a full line of lumped element bandpass, highpass and lowpass filters in low cost TO-style packages. Up to 11 sections are available in highpass and lowpass filters, bandpass filters use up to five sections. Highpass cutoffs as low as 1 MHz are available as are cutoff and center frequencies of up to 2000 MHz for lowpass and bandpass filters.
Lorch Electronics
INFO/CARD #214

TEST EQUIPMENT

Spectrum Analyzers

Anritsu introduces the MS2610 and MS2620 series of spectrum analyzers for testing mobile communications equipment. The MS2612A's frequency range is 9 kHz to 4.6 GHz and the MS2613A's range is from 9 kHz to 6.5 GHz. The MS2621A, MS2622A and MS2623A provide built in tracking generators. The spectrum analyzers are priced at $14,950 for the MS2612A, $17,900 for the MS2613A, $14,780 for the MS2621A, $20,740 for the MS2622A and $22,400 for the MS2623A.
Anritsu Wiltron
INFO/CARD #213

COMINT/SIGINT System

Tecom has developed a 22-pound portable measurement system for 1.7-18 GHz that fits into a standard suitcase. With an

ultra-compact EL/AZ positioner, 12 inch reflector, and linear feed, the unit is designed to create a true manpack COMINT/SIGINT collection system.
Tecom Industries, Inc.
INFO/CARD #212

Noise Figure Meter

The model 8419B noise figure meter is an accurate, rapid, portable unit capable of measuring receiver noise figure from 0 to 25 dB between 10 kHz and 110 GHz, depending on the noise source selection. Model 8419B has a built-in synthesized preselector for suppression of spurious signals and phase-locked IF tuning accuracy.
S.T. Research Corp.
INFO/CARD #211

Waveform Generators

Wavetek has introduced a line of 50 MHz waveform generators that includes the Model 80 function generator, Model 81 pulse/function generator and the Model 85 digitizing arbitrary waveform generator. Model 80 is priced at $2695, Model 81 at $3495 and Model 85 at $3995 with delivery 45 days ARO.
Wavetek Corp.
INFO/CARD #210

Discriminator Delay Line

Oleson Microwave Labs announces the availability of a delay line accessory for use with the HP Model 3048 and other phase noise measurement systems when using discriminator phase noise measurement techniques. Included are a calibrated line stretcher and four separate delay line sections covering 25, 50, 100 and 200 ns. The delay sections can be cascaded for delays totaling up to 375 ns.
Oleson Microwave Labs
INFO/CARD #209

VXIbus Multiplexer

A VXIbus switching module ideal for switching signals up to 200 MHz is now available from Racal-Dana Instruments. The high density multiplexer can be software configured into several matrix dimensions. Model 1260-50B has 80 bi-directional channels, model 1260-50A has 40. Prices for the 1260-50A/B are $2700 and $3700, respectively.
Racal-Dana Instruments, Inc.
INFO/CARD #208

SEMI-CONDUCTORS

PLL Frequency Synthesizer

High speed lock-up, a built-in 1.1 GHz prescaler, power-saving mode, and 2.7 to 5.5 V operation make the HD155001T from Hitachi America an excellent choice for battery powered analog cellular telephones and commercial radio systems. Adjacent channel lock-ups take approximately 2 msec; full range lock-ups require just 25 to 30 msec. The HD155001T is packaged in a

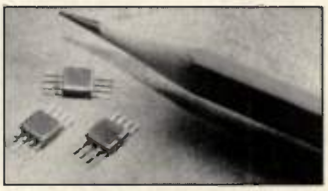
small 20-pin surface mount TSOP package and costs $5.15 in 1000-piece quantities. Samples are currently available.
Hitachi America
INFO/CARD #207

Low Noise Amplifier

Amplifonix announces the model TM5125, a low frequency, low noise figure amplifier. The noise figure is typically 2 dB with a 1.0 dB compression of +24.0 dBm. Third order intercept is typically +40 dBm, and the second order intercept is +52 dBm. VSWR is typically better than 1.7:1 input and 1.35:1 output. TO-8 versions of the unit are available from stock.
Amplifonix, Inc.
INFO/CARD #206

Variable Gain Amplifier

Hewlett-Packard has introduced two new silicon bipolar MMIC variable gain amplifiers, IVA-14228 and IVA-14208. Both devices have operating bandwidths from DC to 2.5 GHz, (3.4 Gb/s data rates). Typical gain is 24 dB, with a 34 dB typical control range (10 dB attenuation can be produced). Both inputs and



outputs can be operated either single ended or balanced. The IVA-14208 is supplied in a plastic SO-8 package and is priced at $9.95 for quantities less than 100. The IVA-14228 is packaged in a high-reliability, metal/ceramic hermetic package and is priced at $54.50 is sub-100 quantities.
Hewlett-Packard Co.
INFO/CARD #205

Low Power Switches

Low leakage (250 pA max. at 25°C), fast switching (175 ns max. turn-on time), and low on-resistance (35 ohms max.) are featured in the DG417, DG418 and DG419 precision CMOS analog switches from Maxim. The DG417 is a SPST normally open switch, the DG418 is a SPST normally closed switch, and the DG419 is a SPDT normally open/normally closed switch. The switches are available in 8-pin DIP and narrow SO packages. Prices start at $1.19 and $1.63 for the single throw and double throw switches, respectively.
Maxim Integrated Products
INFO/CARD #204

Log Amp

By compressing the amplitude of signals, (up to 15 MHz), over a 55 dB range, the TDA8781 logarithmic amplifier from Philips can be used to reduce the resolution needed in A/D converters. Operating from a 5 V supply, it draws a typical current of 8 mA and 250 uA in standby mode. The TDA8781 is packaged in a 14-pin small-outline package.
Philips Semiconductors
INFO/CARD #203

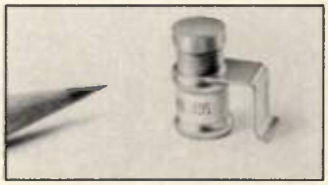
150 MHz Op Amps

SGS-Thomson has announced the TSH150 and TSH151 BiCMOS operational amplifiers. The devices can drive 50 ohm lines directly and have low offset voltages of 5 mV (max), slew rate of 200 V/us, differential gain of 0.05 percent and a differential phase of 0.05 degrees. The THS150 has bipolar input transistors and input noise voltage of 7 nV/$\sqrt{Hz}$. The TSH151 has CMOS input and input bias current of 25 pA (typ). In 10,000 piece quantities, the TSH150 costs $2.60, and the TSH151 costs $3.31.
SGS-Thomson Microelectronics
INFO/CARD #202

DISCRETE COMPONENTS

Vertical Trimmer

Sprague-Goodman has announced the expansion of its Airtrim® air dielectric trimmer capacitor line to include new vertical surface mount models. The new models allow tuning access perpendicular to the mounting surface and are initial-



ly available in three capacitance ranges: 0.35-3.5 pF, 0.6-6 pF and 0.8-10 pF. Model GAA10009 is priced at $6.15 each, in quantities of 100.
Sprague-Goodman Electronics
INFO/CARD #201

Chip Caps

Philips Components' 0603 microwave NP0 ceramic multilayer capacitors exhibit very low insertion loss, even up to 3 GHz. Rated voltage of the series is 63 V and capacitances range from 0.47 pF to 47 pF, with standard tolerances of ±0.5 pF, ±0.25 pF and ±0.1 pF for capacitances below 10 pF.
Philips Components
INFO/CARD #200

Chip Inductors

A new series of discrete chip inductors is being called "the world's smallest" by Toko America because of the series' 1.6 × 0.8mm footprint and 0.8mm height. Self-resonant frequencies are as high as 4 to 5 GHz, and inductances range from 3.9 to 39 nH. Tolerances are 10 or 20 percent, and Q's are over 25 at 300 MHz. The LL1608 series is available on tape and reel in quantities of 4000 pieces per reel.
Toko America, Inc.
INFO/CARD #199

SUBSYSTEMS

IQ Conversion and Processing

The IQ200 is a precision inphase and quadrature (I/Q) signal conversion and filtering system. Two modules are available. The IQ200-RF performs precision I/Q mixing to baseband. The IQ200-LP is a dual channel, programmable lowpass filtering and conditioning system. The two modules together cost $2990, $900 for the IQ200-RF alone, and $2290 for the IQ200-LP alone.
Elanix, Inc.
INFO/CARD #194

Cellular Combiner

The AutoCELL™ autotune cellular Tx combiners use ceramic resonators from Antenna Specialists and technology patented by Decibel Products to speed tuning time and provide automatic frequency re-allocation between cells.
Decibel Products & Antenna Specialists
INFO/CARD #193

Radio Modem

Repco announces the NLR-96, a 9600 bps radio modem requiring no FCC site license. The modem uses spread spectrum modulation and connects to any RS-232 data source to transparently transfer data. It can be configured for either synchronous or asynchronous operation.
Repco, Inc.
INFO/CARD #192

Beamformer

The BM 4422 series of high reliability broadband beamformers combines four input channels and provides four simultaneous outputs with varied amplitudes. Operating from 0.5 to 2.0 GHz, the units meet all performance specs over the –55 to +125 degree C range. Units exhibit 6 dB insertion loss and measure 4.00 × 3.00 × 0.75 inches.
Technical Research and Manufacturing, Inc.
INFO/CARD #190

PC-Controlled, HF Receivers

A group of receivers for stand-alone or multi-channel operation in computer controlled applications has been approved for commercial, industrial and military use. Each unit measures 6.5 × 4.7 × 2 inches, weighs 1.5 pounds and draws less than 10 W of DC power. Frequency coverage is from 5 kHz to 30 MHz with resolution to 1 Hz. Model R232-A costs $5050 and is delivered in 12 weeks.
Inline Components, Inc.
INFO/CARD #189

A Comparison of Simple Software Methods for RF Calculations

By Andrzej Przedpelski
The SHEDD Group

As the old saying goes: "there are many ways of skinning a cat." There are also many ways of performing calculations. The following tutorial is a comparison of some of the possibilities.

There does not seem to be any agreement among engineers as to what method is easiest to perform some of the common design calculations. Four typical methods are presented and their advantages and problems are discussed: BASIC, spreadsheet, math software and a network analysis program. The typical selected examples for these types are: True BASIC v. 3.04, Quattro Pro v. 4.0, MathCAD v. 3.10 and Mike Ellis' RF5. These were selected for the following reasons:

- Comparatively low cost. While "do everything" software will easily perform the most common RF calculations, small companies and individuals usually cannot afford their high cost. The software chosen for this comparison was obtained as lower cost upgrades or during special offer periods.
- Some type of graphic presentation of data was required, since, as Confucius says, "A picture is worth a thousand words."
- The software had to be flexible and suitable for other applications in typical design engineering.

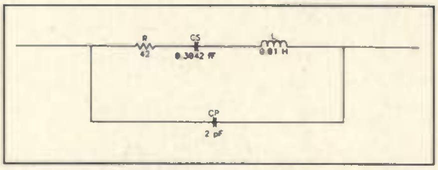


Figure 1. Crystal equivalent circuit.

A typical, though somewhat unusual, design problem was chosen to be solved using these different approaches — Determine the impedance vs. frequency of a high frequency fundamental crystal. This type is not very common and not much data is available. Its series resonance is at approximately 91.2516 MHz and the equivalent circuit, obtained from the manufacturer, is shown in Figure 1. The basic program can be then modified to determine the effect of stray capacitance, resistance, etc.

BASIC Language Programming

The common BASIC software has many drawbacks in engineering work,

```
100 REM CRYSTAL IMPEDANCE CALCULATION
110 REM NOTE: Use LOAD CMPLXLIB (loads complex math) and SCRIPT LOADSG
120 REM       (loads graphics) before running program
130 REM *********List crystal constants*****************************
140 LET Cs=.3042e-15                ! series capacity (Farads)
150 LET Cp=2e-12                    ! parallel capacity (Farads)
160 LET R=42                        ! series resistance (ohms)
170 LET L=0.01                      ! series inductance (Henry)
180 REM **********List frequency range and steps******************
190 LET F1=91.25e6                  ! start frequency
200 LET F2=91.26e6                  ! end frequency (Hz)
210 LET dF=125                      ! frequency steps (Hz)
220 REM **************Start calculation*****************************
230 DIM Mag(81),Ang(81),f(81)
240 LET N=1+(F2-F1)/dF              ! number of points
250 FOR x=1 to N step 1
260     LET f(x)=F1+(x-1)*dF
270     LET XCs=-1/(2*pi*f(x)*Cs)       ! Cs reactance
280     LET XCp=-1/(2*pi*f(x)*Cp)       ! Cp reactance
290     LET XL=2*pi*f(x)*L          ! L reactance
300     LET A$=comp$(42,XCs+XL)     ! complex representation of A
310     LET B$=comp$(0,-1/XCp)      ! complex representation of B
320     LET D$=csum$(cquot$(c_one$,A$),B$)  ! complex D
330     LET Z$=cquot$(c_one$,D$)    ! complex Z(f)
340     LET Mag(x)=cabs(Z$)         ! magnitude
350     LET Ang(x)=cang(Z$)*360/(2*pi)      ! phase angle in degrees
360     PRINT f(x),Mag(x),Ang(x)    !  print results
370 NEXT x
380 REM ***************Plotting routines*********************************
390 CALL magnitude
400 CALL angle
410 REM *****************Magnitude plot**********************************
420 SUB magnitude
430     OPEN #1: screen 0,.49,.1,.9     ! left half size window
440     CALL SetGraphType("LOGY")       ! y-axis logarithmic
450     CALL SetHlabel ("Freq, Hz")     ! label horizontal axis
460     CALL SetVlabel ("Magnitude")    ! label vertical axis
470     CALL SetTitle ("MAGNITUDE")     ! label graph
480     CALL SetGrid ("H.V.")       ! grid style
490     CALL SetXscale (F1,F2)      ! puts limits on X scale
500     CALL SetYscale (10,100000)      ! puts limits on Y scale
510     CALL DataGraph (f,Mag,0,1,"red white red")     ! plots f and M
520 END SUB
530 SUB angle
540     OPEN #2: screen .51,1,.1,.9     ! right half window
550     CALL SetGraphType ("XY")
560     CALL SetHlabel ("Freq, Hz")
570     CALL SetVlabel ("Angle, deg")
580     CALL SetTitle ("PHASE ANGLE")
590     CALL SetGrid ("H.V.")
600     CALL SetXscale (F1,F2)
610     CALL SetYscale (-90,-90)
620     CALL DataGraph (f,Ang,0,1, "red white red")
630 END SUB
640 END
```

Table 1. BASIC program for calculating crystal impedance.

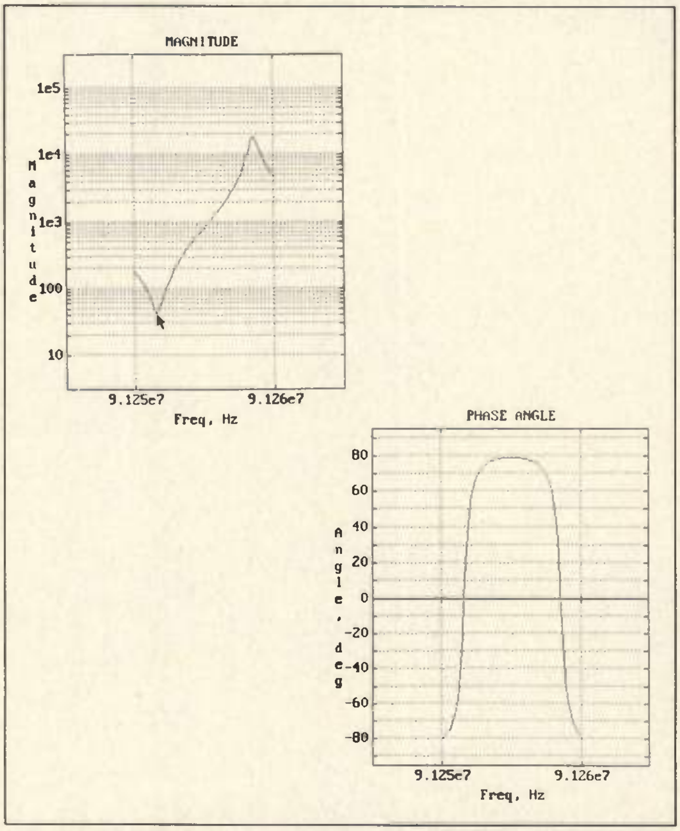


Figure 2. Magnitude and phase plots generated with a BASIC program.

RF TRANS

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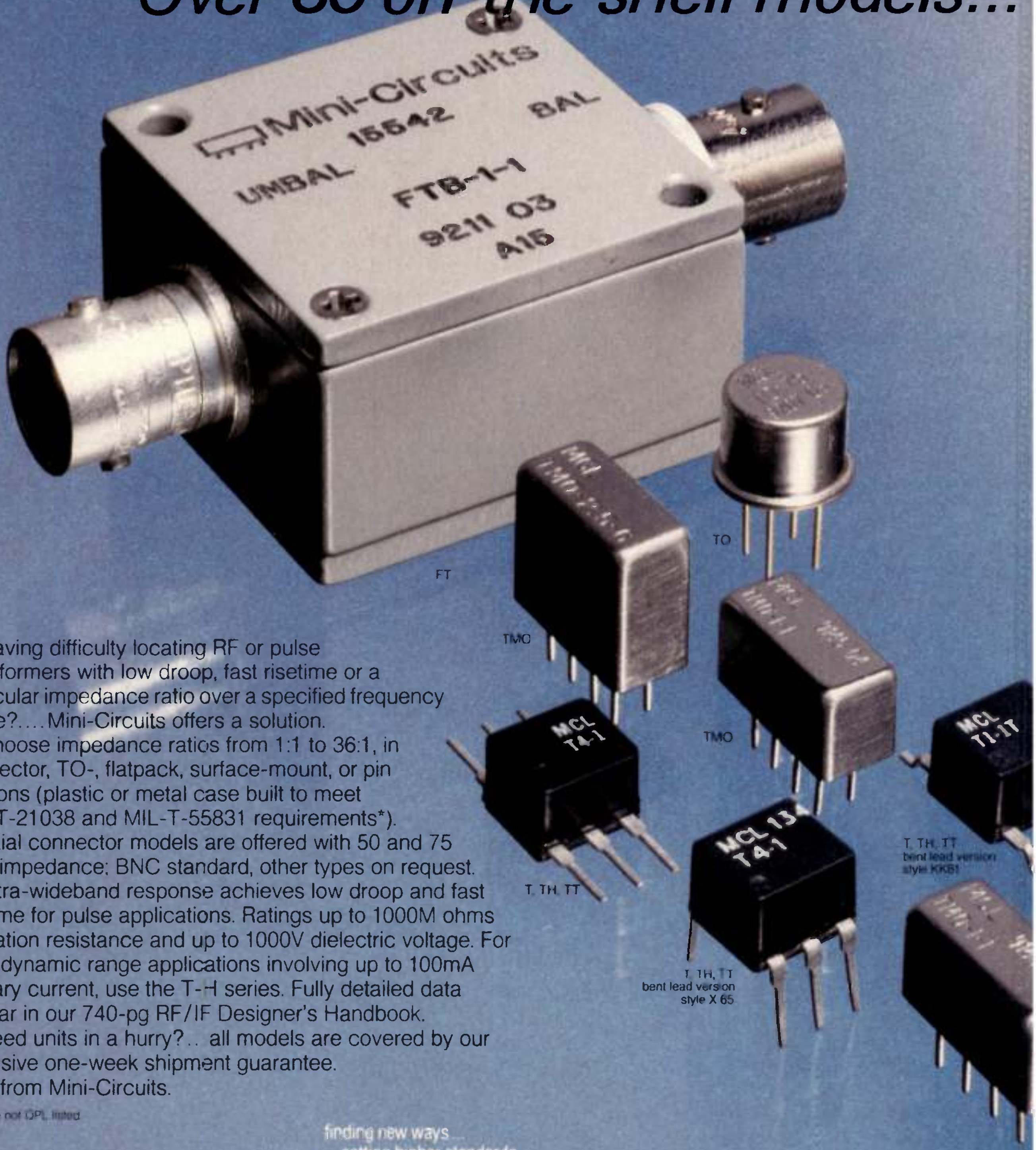

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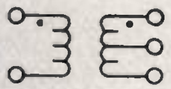
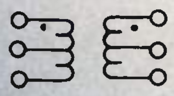
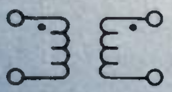
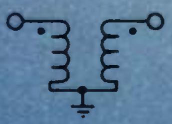
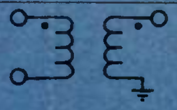
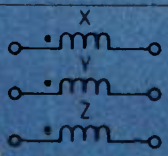
		MODEL NO.	Ω RATIO	FREQUENCY MHz	INSERTION LOSS 3dB MHz	INSERTION LOSS 2dB MHz	INSERTION LOSS 1dB MHz	PRICE $ Qty. (1-9)
A* 	T	T1-1T	1	.05-200	.05-200	.08-150	.2-80	4.45
		T1-6T	1	.015-300	.015-300	.021-150	.03-50	6.95
		T2-1T	2	.07-200	.07-200	.1-100	.5-50	4.95
		T2.5-6T	2.5	.01-100	.01-100	.02-50	.50-20	4.95
		T3-1T	3	05-250	.05-200	.1-200	.5-70	4.95
		T4-1	4	.2-350	.2-350	.35-300	2-100	3.25
		T4-6T	4	02-250	.02-250	.05-150	0.1-100	4.45
		T5-1T	5	.3-300	.3-300	.6-200	5-100	4.95
		T8-1T	8	0.3-140	0.3-140	0.7-90	1-60	7.95
		T13-1T	13	.3-120	.3-120	.7-80	5-20	4.95
		T16-6T	16	.03-75	03-75	06-30	.1-20	5.65
	TH	T4-1H	4	10-350	10-350	15-300	25-200	5.95
	TMO	TMO1-1T	1	.05-200	05-200	.08-150	2-80	7.95
		TMO2-1T	2	.07-200	.07-200	1-100	.5-50	8.45
		†TMO2.5-6T	2.5	.01-100	.01-100	.02-50	.05-20	8.45
		†TMO3-1T	3	.05-250	.05-250	.1-200	5-70	7.95
		TMO4-1	4	.2-350	.2-350	.35-300	2-100	6.25
		TMO5-1T	5	3-300	.3-300	6-200	5-100	8.45
		TMO13-1T	13	3-120	.3-120	.7-80	5-20	8.45
B* 	TT	TT1-6	1	004-500	.004-500	.02-200	.1-50	6.95
		TT1.5-1	1.5	075-500	075-500	.2-100	1-50	5.95
		TT2.5-6	2.5	.01-50	.01-50	.025-25	.05-10	6.45
		TT4-1	3	.05-200	.2-50	.2-50	1-30	5.95
		TT4-1A	4	0.1-300	0.1-300	0.2-250	0.3-180	6.95
		TT25-1	25	.02-30	.02-30	.05-20	.1-10	9.95
	TTMO	TTMO25-1	25	02-30	.02-30	05-20	.1-10	11.95
		TTMO1-1	1	005 100	.005-100	.01-75	.05-40	11.45
		TTMO4-1A	4	0.1-300	0.1-300	0.2-250	0.3-180	13.95
C 	T	T1-1	1	.15-400	.15-400	.35-200	2-50	3.25
		T1.18-3	1.18	0.01-250	0.01-250	0.02-200	0.03-50	5.65
		T1-6	1	.01-150	01-150	02-100	.05-50	5.65
		T1.5-1	1.5	.1-300	.1-300	.2-150	5-80	4.45
		T1.5-6	1.5	.02-100	02-100	.05-50	0.1-25	5.65
		T2.5-6	2.5	01-100	01-100	02-50	.05-20	4.45
		T4-6	4	.02-200	.02-200	.05-150	1-100	4.45
		T9-1	9	.15-200	.15-200	.3-150	2-40	3.95
		T16-1	16	.3-120	3-120	.7-80	5-20	4.45
		T36-1	36	.03-20	.03-20	.05-10	1-5	6.95
	TO	TO-75	1	10-500	—	10-500	40-250	6.95
	TH	T1-1H	1	8-300	8-300	10-200	25-100	5.95
		T9-1H	9	2-90	2-90	3-75	6-50	6.45
		T16-1H	16	7-85	7-85	10-65	15-40	6.45
	TMO	TMO1-02	1	1-800	1-800	2-500	—	9.45
		TMO1-1	1	.15-400	.15-400	.35-200	2-50	6.25
		TMO1.5-1	1.5	.1-300	.1-300	2-150	.5-8	8.45
		†TMO2.5-6	2.5	.01-100	.01-100	.02-50	05-20	7.95
		†TMO4-6	4	02-200	02-200	.05-150	1-100	7.95
		TMO6-1	6	3-200	.3-200	.5-150	5-50	7.95
		TMO9-1	9	15-200	.15-200	3-150	2-40	7.95
		TMO16-1	16	3-120	.3-120	.7-80	5-20	7.95
D 	T	T2-1	2	050-600	.050-600	1-400	5-200	3.95
		T3-1	3	5-800	5-800	2-400	—	4.45
		T4-2	4	2-600	.2-600	5-500	2-250	3.95
		T8-1	8	.15-250	15-250	25-200	2-100	3.95
		T14-1	14	2-150	2-150	5-100	2-50	4.95
	TMO	TMO2-1	2	050-600	050-600	1-400	5-200	7.95
		TMO3-1	3	.5-800	5-800	2-400	—	8.45
		TMO4-2	4	.2-600	2-600	5-500	2-250	7.95
		TMO8-1	8	15-250	15-250	25-200	2-100	7.95
		TMO14-1	14	2-150	2-150	5-100	2-50	8.45
	FT	FT1.22-1	1.22	005-100	005-100	01-50	05-25	35.95
		FT1.5-1	1.5	.1-400	1-400	5-200	1-100	35.95
E 	FTB	FTB-1	1	2-500	2-500	5-300	1-100	36.95
		FTB1-6	1	01-125	01-125	05-50	1-25	36.95
		■FTB-1-75	1	5-500	5-500	5-300	10-100	36.95
F 	T	T-622	1	01-200	01-200	05-100	5-80	3.25
		T626	1	0.01-10	0.01-10	0.2-5	04-2	3.95
		T2-5131						

■ Denotes 75 ohm models

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Typical Amplitude Unbalance
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NEW TC SURFACE MOUNT MODELS from 1 MHz to 1500 MHz

INFO/CARD 53

F 72 REV. D

CRYSTAL IMPEDANCE

R=	42
L=	0.01
CS=	3.04E-16
CP=	2E-12
F1=	91250000
F2=	91260000
DF=	250

FREQ	XCS	XL	XCP	XL+XCS	A	angA	B	C	D	MAG	ANG
91250000	-5733609	5733407	-872.082	-202	0.0048	1.366	0.00099	0.00474	0.0059	167.44	-80.50
91250250	-5733593	5733422	-872.079	-17·	0.0057	1.330	0.00136	0.00552	0.0067	146.87	-78.48
91250500	-5733577	5733438	-872.077	-139	0.0069	1.278	0.00199	0.00658	0.0077	125.31	-75.59
91250750	-5733562	5733454	-872.075	-108	0.0086	1.199	0.00314	0.00805	0.0092	102.90	-71.17
91251000	-5733546	5733469	-872.072	-76	0.0115	1.068	0.00552	0.01005	0.0112	80.10	-63.74
91251250	-5733530	5733485	-872.070	-45	0.0162	0.820	0.01108	0.01188	0.0130	58.48	-49.60
91251500	-5733514	5733501	-872.068	-14	0.0227	0.313	0.02155	0.00697	0.0081	43.42	-20.64
91251750	-5733499	5733517	-872.065	18	0.0219	-0.401	0.02017	-0.00856	-0.0074	46.52	20.19
91252000	-5733483	5733532	-872.063	49	0.0155	-0.865	0.01003	-0.01176	-0.0106	68.51	46.62
91252250	-5733467	5733548	-872.060	81	0.0110	-1.091	0.00508	-0.00975	-0.0086	100.07	59.46
91252500	-5733452	5733564	-872.058	112	0.0084	-1.212	0.00293	-0.00782	-0.0067	137.13	66.29
91252750	-5733436	5733579	-872.056	143	0.0067	-1.286	0.00188	-0.00642	-0.0053	178.66	70.39
91253000	-5733420	5733595	-872.053	175	0.0056	-1.335	0.00130	-0.00541	-0.0043	224.60	73.05
91253250	-5733404	5733611	-872.051	206	0.0047	-1.370	0.00095	-0.00465	-0.0035	275.26	74.88
91253500	-5733389	5733627	-872.048	238	0.0041	-1.396	0.00072	-0.00406	-0.0029	331.18	76.19
91253750	-5733373	5733642	-872.046	269	0.0037	-1.416	0.00057	-0.00363	-0.0025	393.07	77.15
91254000	-5733357	5733658	-872.044	301	0.0033	-1.432	0.00046	-0.00326	-0.0021	461.86	77.84
91254250	-5733342	5733674	-872.041	332	0.0030	-1.445	0.00038	-0.00296	-0.0018	538.70	78.34
91254500	-5733326	5733689	-872.039	363	0.0027	-1.456	0.00031	-0.00272	-0.0016	625.05	78.69
91254750	-5733310	5733705	-872.036	395	0.0025	-1.465	0.00027	-0.00250	-0.0014	722.72	78.90
91255000	-5733295	5733721	-872.034	426	0.0023	-1.473	0.00023	-0.00232	-0.0012	834.06	78.99
91255250	-5733279	5733736	-872.032	458	0.0022	-1.479	0.00020	-0.00217	-0.0010	962.17	78.97
91255500	-5733263	5733752	-872.029	489	0.0020	-1.485	0.00017	-0.00203	0.0009	1111.02	78.83
91255750	-5733247	5733768	-872.027	520	0.0019	-1.490	0.00015	-0.00191	0.0008	1286.07	78.57
91256000	-5733232	5733784	-872.025	552	0.0018	-1.495	0.00014	-0.00180	-0.0007	1494.81	78.17
91256250	-5733216	5733799	-872.022	583	0.0017	-1.499	0.00012	-0.00171	-0.0006	1747.89	77.60
91256500	-5733200	5733815	-872.020	615	0.0016	-1.503	0.00011	-0.00162	-0.0005	2060.87	76.82
91256750	-5733185	5733831	-872.017	643	0.0015	-1.506	0.00010	-0.00154	-0.0004	2457.45	75.75
91257000	-5733169	5733846	-872.015	673	0.0015	-1.509	0.00009	-0.00147	-0.0003	2975.33	74.27
91257250	-5733153	5733862	-872.013	709	0.0014	-1.512	0.00008	-0.00141	-0.0003	3678.04	72.17
91257500	-5733137	5733878	-872.010	743	0.0013	-1.514	0.00008	-0.00135	-0.0002	4679.83	69.08
91257750	-5733122	5733894	-872.008	772	0.0013	-1.516	0.00007	-0.00129	-0.0001	6202.34	64.15
91258000	-5733106	5733909	-872.005	808	0.0012	-1.519	0.00006	-0.00124	-0.0001	8700.19	55.61
91258250	-5733090	5733925	-872.003	835	0.0012	-1.521	0.00006	-0.00120	-0.0000	12958.73	38.80
91258500	-5733075	5733941	-872.001	866	0.0012	-1.522	0.00006	-0.00115	-0.0000	17822.01	5.32
91258750	-5733059	5733956	-871.998	897	0.0011	-1.524	0.00005	-0.00111	0.0000	15954.29	-33.88
91259000	-5733043	5733972	-871.996	929	0.0011	-1.526	0.00005	-0.00107	0.0001	11470.06	-56.14
91259250	-5733028	5733988	-871.993	960	0.0010	-1.527	0.00005	-0.00104	0.0001	8573.99	-67.06
91259500	-5733012	5734003	-871.991	992	0.0010	-1.528	0.00004	-0.00101	0.0001	6823.48	-73.09
91259750	-5732996	5734019	-871.989	1023	0.0010	-1.530	0.00004	-0.00098	0.0002	5693.21	-76.82
91260000	-5732980	5734035	-871.986	1055	0.0009	-1.531	0.00004	-0.00095	0.0002	4913.50	-79.32

Table 2. Spreadsheet for calculating crystal impedance.

including lack of π, log10, complex number calculations and graphics. My first experience with BASIC was using an HP-86 with the built-in HP BASIC, which had provisions for these functions. Thus, I may be somewhat spoiled. After switching to a "compatible" computer, I had to look for some other solutions. Very suitable BASIC software is available from TransEra and HP, but is somewhat expensive and did more than I needed. I finally settled on True BASIC (1). One of its advantages, in addition to the low cost, is the possibility of tailoring it to your needs by using the "Toolkit" additional software. For my work I chose the True BASIC Language System with the Mathematician's Toolkit which, among other functions, includes complex number calculations, and the Scientific Graphics Toolkit, which lets you chose among several types of graphs and facilitates printing. [*It doesn't really matter what your favorite programming language is. Most of these comments about BASIC apply to writing your own programs in other popular languages, such as Fortran, C or Pascal — Editor*]

The first step is to reduce the problem to a formula, which can be then solved using a BASIC program. The second step is to write this program, run it, debug it (if necessary) and then print the graph.

Some engineers argue that you can buy programs which do all this for you and all you have to do is enter the data. I don't necessarily agree with this for two reasons: Deriving the formula and writing your own program gives you an idea of how the circuit works and how it can be possibly modified; and, if something goes wrong you may be able to adjust the program to rectify the problem. I have had cases, in unusual circuits (and who dces not run into these?) where the "canned" programs would give me obviously wrong answers. This can get you into trouble, since some engineers believe the computer explicitly (I learned the hard way to be more careful)!

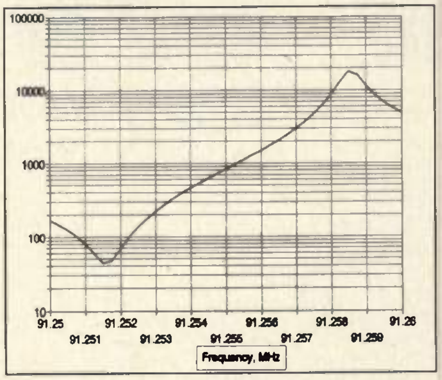


Figure 3. Magnitude plot generated by Quattro Pro.

To begin the analysis of the crystal impedance, we use:

$$X_L = j2\pi fL \tag{1}$$

$$X_C = 1/j2\pi fC \tag{2}$$

where L is the inductance in Henries (0.01), C is the capacitance in Farads (0.3042×10^{-15} for the series capacitance and 2×10^{-12} for the parallel capacitance) and the series R (42 ohms). We obtain, using the series and parallel impedance combining, the overall impedance formula, as a function of frequency:

$$\left[\left(42 + j2\pi f \cdot 0.01 + \left(j2\pi f \cdot 0.342\times10^{-15}\right)^{-1}\right)^{-1} + \left(j2\pi f \cdot 2\times10^{-12}\right)^{-1}\right]^{-1} \tag{3}$$

We can now write the BASIC program

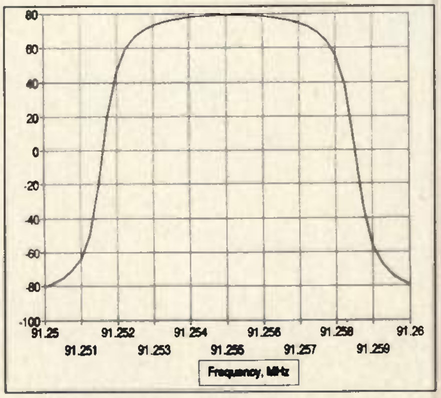


Figure 4. Phase plot generated by Quattro Pro.

to obtain the desired graph. I like to break up the program into smaller pieces and to add comments and to make debugging and modifying easier:

$$D \text{ (denominator)} = 1/A + B, \text{ and} \quad (4)$$

$$Z \text{ (impedance)} = 1/D \quad (5)$$

The BASIC program is shown in Table 1 and the resulting graphs in Figure 2. Log Y-axis is used for the magnitude to show better the series and parallel resonance points because of the large differences in values. Line 360 prints the data and can be deleted, if not necessary. The two graphs are displayed side-by-side to show their relationships. However, they can be made full page size by changing 430 and 540 to 0,1,0,1. As a result, more labeled frequency grids are shown, making reading easier.

Spreadsheet

To use the spreadsheet, the calculation can be broken into steps as shown below:

$$\frac{1}{A\angle angA + j\frac{1}{XCP}} \quad (6)$$

$$\frac{1}{B + jC + j\frac{1}{XCP}} \quad (7)$$

$$\frac{1}{B + j\left(C + \frac{1}{XCP}\right)} = \frac{1}{B + jD} \quad (8)$$

A typical spreadsheed, using Quattro Pro (2), is shown in Table 2. The first column is the frequency. It can be set-up using the "FILL" command. The next 3 columns show the reactances of the components at the given frequency. The next 6 columns show the computation as it proceeds. Finally, the last 2 columns give the desired results: the magnitude and the phase of the impedance. Having the in-between calculations simplifies debugging and also gives a better insight into what is happening. The XL+XCS column shows how the series reactance changes from capacitive to inductive at the series resonance point. The next 5 columns are defined as follows:

$$A = \left|1/\left(R(XL + XCS)\right)\right| \quad (9)$$

$$angA = -\text{atan}[(XL+XCS)/R] \quad (10)$$
(in radians)

$$B = \text{real part of } A \quad (11)$$

$$C = \text{imaginary part of } A \quad (12)$$

$$D = C + 1/XCP \quad (13)$$

The last two, giving the answers, are:

$$MAG = \frac{1}{|B + D|} \quad (14)$$

$$ANG = -\text{atan}(D/B) \text{ (in degrees)} \quad (15)$$

The graphs, shown in Figures 3 and 4, were made using the XY graph format and use the frequency column for the X-axis and the MAG and ANG columns respectively for the Y-axis. If desired, they can be combined into a single graph.

To reduce the size of Table 2, 250 Hz steps (instead of 125 Hz) were used. This produced the somewhat lumpy Figures 3 and 4 curves at resonance points. For best results, the smaller steps should be used for the final graphs.

Math Software

Using Mathcad (3), the process is comparatively simple, as shown in Figure 5. The frequency range and step is first specified by giving the start, the second frequency and the end frequencies. Then, the formulas are written. First the basic impedance formula (same as equation 3), then the magnitude and the angle of this function. These are shown with the (:=) equal signs indicating that they are definitions. If a printout of the data is desired, the headings f=, M(f)= and A(f)= will give three columns of data. This time the (=) sign is used indicating that an answer is required. Using the GRAPH functions, the desired graphs are obtained. A lot of options are available: text font, constants font, variables font, graph format, etc. While comparatively easy to use, some caution is required. If the answers are 0, for instance, this is an indication that the answer is less than the default zero value. This can be corrected by changing this default value. If the answers are obviously wrong, as when calculating a differential of an unusual function, the value of the default tolerance may have to be changed.

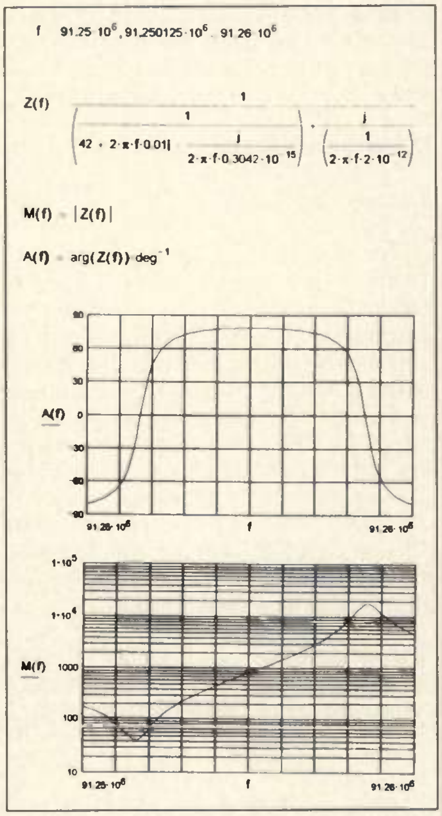


Figure 5. Magnitude and phase calculated and plotted on a page of MathCAD.

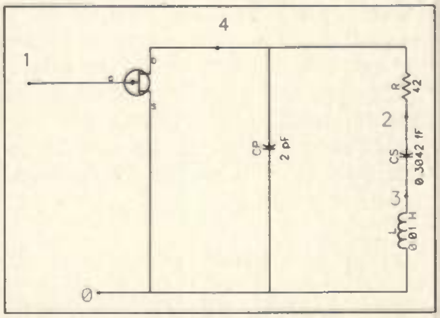
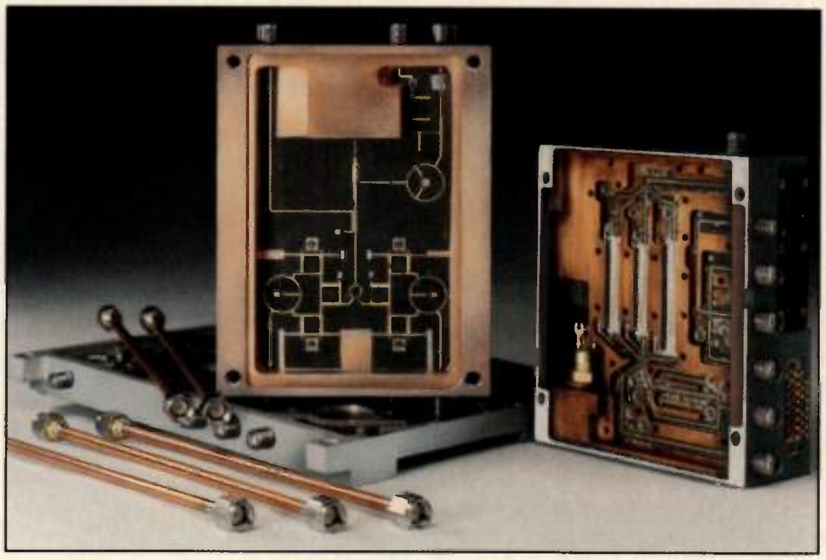
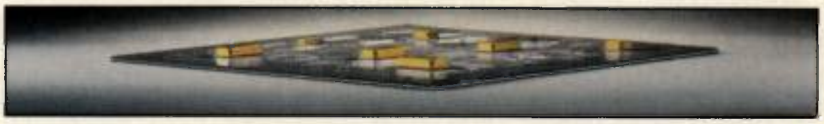


Figure 6. Network used for impedance calculations using RF5 circuit analysis program.

Now you can fit all of this



into this.



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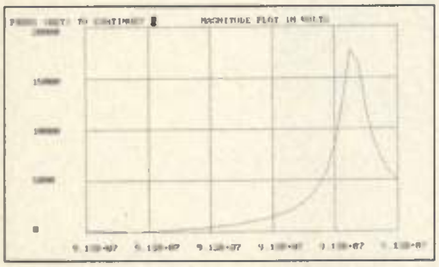

Figure 7. Magnitude plot using linearly scaled y-axis. Additional signifigant figures must be added to the x-axis by hand.

Network Program

A nodal network analysis program can also be used to obtain the desired answers. A program such as Gary Appel's ACANAL (4) would be suitable for most applications. However, in this case it cannot be used since the program allowed minimum capacitance value is 0.01 pF and the value of the series capacitor is less than that. A program such as LCAS provides good data in table form, but the graphs are confusing since the Y-axis reference is offset by an odd number making graph reading difficult. Michael Ellis' program RF5 (5) is used here as an example. This program does not calculate impedance directly, as is the case with some other similar programs. To get around this problem, a constant current source (using the available FET) is used. Since:

$$Z = V/I \qquad (16)$$

and gain of an idealized FET is

$$I/V_{in} = g_m \qquad (17)$$

then

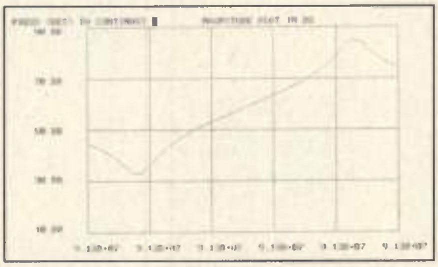

Figure 8. Impedance magnitude plotted in dB.

$$Z = V/g_m \times V_{in} \qquad (18)$$

where V = the output of the FET.

Using the above, the circuit of Figure 6 can be used if we use $g_m = -1$ (the FET normally inverts the phase). Now, the gain from node 1 to node 4 will give us the desired Z. Two more problems remain. The program is set up for labeling the X-axis with only 3 significant figures, and 5 or 6 are needed. Thus, the labeling has to be changed on the graph by hand. The other problem involves the very large differences between the impedances at series and parallel resonances. Using a single Y-axis scale, as in Figure 7, the series resonance is not readable. One solution is to break up the total frequency range into two parts: one covering the series resonance and the other parallel resonance. This is not a very elegant solution. A better one would be to use a log Y-axis. Since it is not available in this program the gain in dB plot can be used, as shown in Figure 8. While expressing impedance in dB may not be common, it provides the desired results. The Y-axis scale can then be converted using dB = 20 log (Z)

This new scale can then be substituted manually. The phase, using this program, is plotted in Figure 9. Both curves are not very smooth at the resonance points, since the program, as written, did not allow the use of 125 Hz steps (250 Hz steps were used).

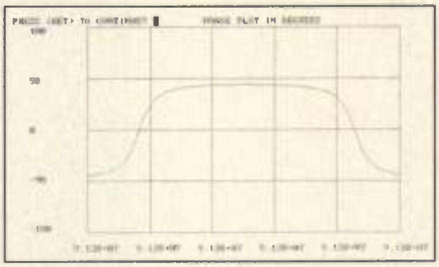

Figure 9. Impedance phase plot.

Conclusions

A few alternative methods for computer analysis of a simple circuit have been presented. The above example is somewhat unusual because of the sharp resonances, but even so, several different techniques can be used to obtain the desired data. If the source code is available for the low cost or public domain programs, they can be modified to provide a more desirable format of the required outputs. If not, some other "fixes" are possible, as shown in the Network Program case. The above examples are by no means all of the possible solutions. For example, in some cases, Smith chart programs can be used. *RF*

Notes and References

1. True BASIC, Inc., West Lebanon, NH.
2. Quattro Pro is from Borland International, Scotts Valley, CA.
3. MathCAD is a product of MathSoft, Inc., Cambridge, MA.
4. G. Appel, "A Nodal Network Analysis Program," *RF Design*, November 1989, p. 82. Available from the RF Design Software Service as disk number RFD-1189.
5. M. Ellis, "A Comprehensive Filter Design Program," *RF Design*, July 1991, p. 31. RF5 is the circuit analysis module of this program, which may be used as a stand-alone. The program is number RFD-0791 from the RF Design Software Service.

About the Author

Andrzej (Andy) Przedpelski is the consulting editor for *RF Design* and founder of the SHEDD Group, an RF design, military product consulting group. He can be reached at 7260 Terrace Place, Boulder, CO 80303-4638, or by phone at (303) 499-9517.

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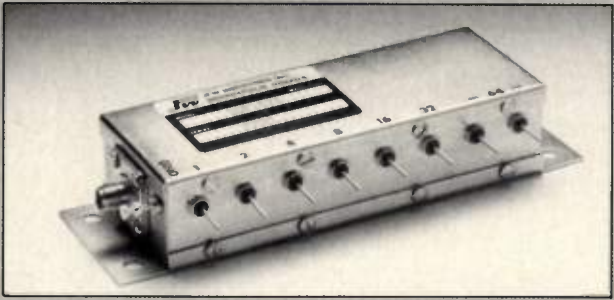

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Attenuation Range
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Frequency Range
50-250 MHz
Attenuation Range
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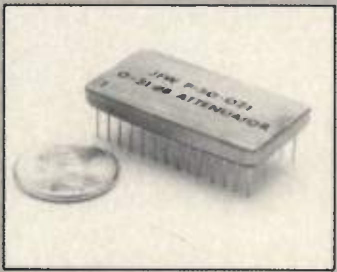

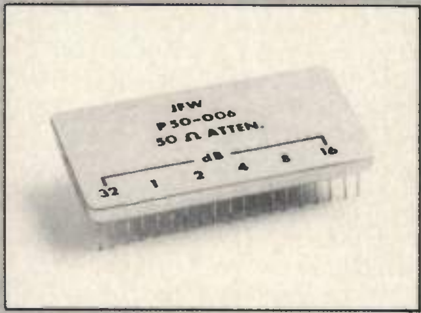

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INFO/CARD 106

A Program for Impedance Format Conversion

By Thomas Bavis
ENI

This impedance conversion program was written as a way to relate measurement results from a variety of equipment, each having a different output format. Like many single-purpose tools, it is much faster to use than a general-purpose one. It is still used frequently, eight years after it was written in Applesoft BASIC.

Using the program is simple. Type Z and hit enter to start. A list of impedance formats is shown — series or parallel R and X, polar Z and θ, reflection coefficient, SWR, etc. Enter a system impedance (or hit enter for the default value of 50 ohms), then press enter or an arrow key until the cursor is on the format you wish to enter.

After entering the value, the remaining formats are calculated and displayed. Scalar entries (VSWR, return loss, etc.) are converted only to other scalar formats. Press Q to quit, or enter another value as you did before. Figures 1 and 2 show examples that demonstrate the program's ability to convert scalar and complex impedances.

These are the formulas needed for the impedance format conversions performed in this program. (the job of inverting the equations to do the conversions in reverse is left to the reader.)

$R_o \equiv$ system impedance

$\rho \equiv$ reflection coefficient

RL ≡ return loss (dB)

s subscripts indicate series elements

p subscripts indicate parallel elements

$$R_p = \frac{R_s^2 + X_s^2}{R_s}, \quad X_p = \frac{R_s^2 + X_s^2}{X_s}$$

$$\rho = \sqrt{\frac{(R_s - R_o)^2 + X_s^2}{(R_s + R_o)^2 + X_s^2}} = 10^{-\frac{RL}{20}}$$

$$\text{polar:} \quad Z = \sqrt{R_s^2 + X_s^2}$$

$$\theta = \operatorname{atan} \frac{X_s}{R_s}$$

$$\text{reflected power } (\%) = 100\rho^2$$

It takes less time for you to run the program than it would to look up a formula or table. Formulas used in the program came from the *Radio Amateur's Handbook* from the Amateur Radio Relay League (ARRL).

Conversion to other BASIC dialects should not be difficult — It was converted to GWBASIC in 20 minutes by transferring it via serial cable, then using a word processor to search and replace the APPLE specific keywords (HTAB to LOCATE, etc.)

This program is available on disk, (in GWBASIC and in compiled form), through the RF Design Software Service. See page 84 for ordering information. ***RF***

About the Author
Thomas Bavis has worked in product support and new product development at ENI since 1989. Prior to his current job, he worked at Scientific Radio Systems. He can be reached at ENI at 100 Highpower Rd., Rochester, NY 14623, or by phone at (716) 292-7532.

```
IMPEDANCE CALCULATION PROGRAM

<Q> : Quit       <Any> : Continue

SYSTEM IMPEDANCE : 50

SERIES R,X :10,12.5 Normalized :.2, .25

PARALLEL R, X : 25.63 , 20.5

POLAR Z, THETA : 16.01 , 51.34 Deg

REFL COEFF : .684

SWR : 5.32 : 1

% REFL PWR : 46.76 %

RETURN LOSS : 3.3 dB
```

Figure 1. Conversion of a complex impedance.

```
IMPEDANCE CALCULATION PROGRAM

<Q> : Quit       <Any> : Continue

SYSTEM IMPEDANCE :  50

SERIES R, X :

PARALLEL R, X :

POLAR Z, THETA :

REFL COEFF :  .091

SWR : 1.2 : 1

% REFL PWR : .83 %

RETURN LOSS : 20.83 dB
```

Figure 2. Conversion of a scalar impedance.

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INFO/CARD 58

F163 REV ORIG

Continuously Variable Bandwidth Feedback Demodulator for Home FM Stereo Reception

By Jerry Pulice
Smith's Industries

In the Princeton, New Jersey area, halfway between New York City and Philadelphia, every FM broadcast channel, (spaced 200 kHz from 88 tc 108 MHz in North America) is occupied by a transmitter with high received signal strength. During FM stereo reception, interference from adjacent channel transmitters presents a severe problem. Even with a rotatable, highly directional outdoor antenna it is difficult to achieve high quality, noise free FM stereo reception. This is because the requirements for good adjacent channel selectivity compete with the requirements for clean stereo FM reception.

In this paper I present a simple, very low cost method to easily configure the home receiver for any compromise between low distortion using wide IF band-width and greater adjacent channel rejection using a high selectivity IF bandwidth.

Review of FM Theory and Practice

Carson's Rule (1) gives a good working number for the bandwidth occupied by an FM transmitter: twice the sum of the peak carrier deviation and the highest modulation frequency present. In the case of monaural broadcast, this would be 2 × (75 kHz deviation plus 15 kHz modulation limit) or 180 kHz (±90 kHz). This generally gives the spectrum occupied by 99 percent of the transmitter's energy, for most typical modulation inputs. With a channel spacing of only 200 kHz (±100 kHz), any receiver intended to achieve good adjacent channel rejection would require IF filters with extreme shape factors.

FM IF filters are designed for phase linearity (constant Δphase/Δfreq, or constant group delay) across their passband (2) and, in general, such filters do not have idealized rectangular passbands. This leads to degraded adjacent channel rejection.

In 1961, a suppressed carrier AM-on-FM stereo multiplex subchannel was added on as an afterthought: the increased bandwidth requirements made the above problems worse. This stereo subcarrier added baseband frequencies up to 53 kHz. The subcarrier, (but not the added 19 kHz pilot tone used for sync), tends to vanish when the signal tends to mono (little difference between left and right channels), and 75 kHz peak carrier deviation is permitted at this time. Large amounts of stereo information (left or right channel predominant) reduce the peak carrier deviation permitted to only 45 percent of the mono spec in an effort to keep down the need for bandwidth (3). Even so, the new format needs even more bandwidth than before. Typical ceramic filters sold for FM stereo use have bandwidths from 230 to 280 kHz (4).

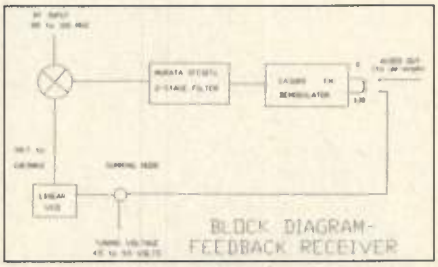

Figure 1. System block diagram.

Since 200 kHz channel spacing was maintained, adjacent channel receiver performance was degraded further. Other subcarriers, as high as 91 kHz, are also used for such services as SCA (background music services, data, etc.)

FM IF Bandpass Filter Practice

The classical IF filter design best suited for flat group delay FM service is the Bessel characteristic — noted for a slow rate of rolloff and poor shape factor. Other classical filter designs have "dog eared" phase response and can be compensated for flat group delay using all-pass phase compensation networks which can have flat amplitude response, but specified phase response. In the complex plane such filters have right hand plane zeros, and their transfer functions can be visualized using the method taught in school using the length and angle from the origin to each pole and zero. Compensated LC filters using this technique were used in the McIntosh MR-78 tuner, but this was an extremely expensive audiophile-targeted design not representative of the average FM receiver. Articles on compensated crystal filters have been published in *RF Design* (5).

Modern SAW filters are non-minimum phase designs which can specify phase and amplitude response independently and would seem to be the ideal candidate for a modern FM receiver IF filter. The very narrow FM radio bandwidth (relative to TV IF which uses SAW filters) and/or the very low 10.7 MHz IF traditionally used in FM receivers lead to large amounts of expensive SAW filter substrate which is probably why low cost SAW filters for commercial FM reception have never been developed.

Current IF Design Practice for High Quality FM Stereo

To accommodate the wide range of reception conditions usually present, high end manufacturers such as Sequerra and McIntosh provide their top-of-the-line receivers with three selectable phase-linear IF filters as follows:

One of uncompromising bandwidth (of

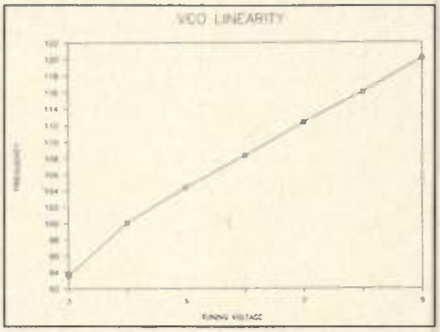

Figure 2. Measured VCO linearity.

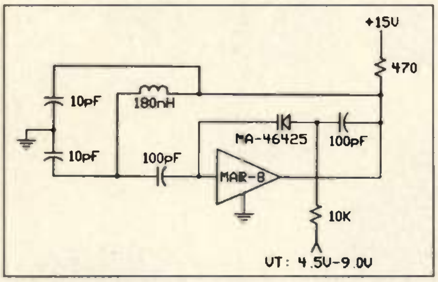

Figure 3. Linear VCO schematic — output 98.7 to 118.7 MHz.

about 280 kHz) for strong signals not corrupted by adjacent channel splatter, and where the highest audio quality is wanted. The adjacent channel(s) must be essentially vacant when receiving with this filter.

Another of minimal bandwidth (about 180 kHz) where acceptable monophonic reception is about all that can be hoped for in the presence of strong adjacent channel splatter, and where the highest audio quality is wanted. Some stereo reception is possible, but with an increase in distrortion over the wideband filter above.

Finally, a compromise filter with bandwidth of about 230 kHz, about equal to that used in the average home FM stereo receiver, to be used for average reception conditions and with average results.

Clearly the above choices are not an economic solution for anything less than the most expensive receiver designs, so an experiment was tried where only one high quality, group delay equalized filter is used only during the most difficult reception conditions. Special means are used to "shoehorn" wider received signals through it when conditions warrant, resulting in decreased distortion and improved stereo decoding. Such means, in general, rely on placing feedback around the demodulator to instantaneously reduce the deviation of the FM signal and were first described by H.C. Chaffee in 1939 (6).

FM Demodulators Currently Employing Negative Feedback

Feedback FM demodulators are actually now in common use in one consumer application. Designers of home TVRO systems operating in the 3.7 to 4.2 GHz downlink band work under severe cost constraint, since these systems are intended for backyard use. Textbook FM demodulators typically require a 10 dB carrier-to-noise ratio in their IF bandwidth to achieve "threshold"; below this CNR, demodulated

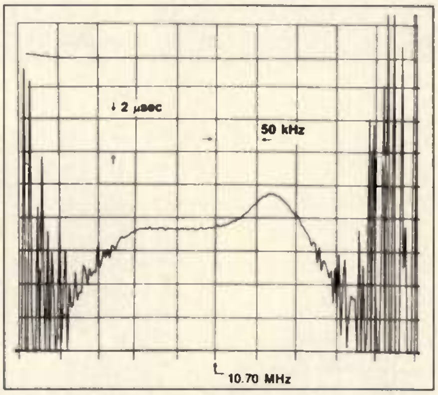

Figure 4a. Group delay vs. frequency for two cascaded Murata SFA 107z filters.

video output quality degrades so rapidly the system becomes useless (7). Methods currently in use to overcome this threshold effect are one example of the above mentioned feedback technique.

Threshold Extension

A three-dB increase in threshold performance is equivalent to doubling the area of the receiving dish, therefore, it is not surprising that TVRO designers incorporate this "threshold extension" into their receivers. A ten foot dish then becomes comparable to a fourteen foot dish when operated near FM threshold by incorporating this feature in the receiver electronics.

One way to achieve threshold extension in an FM system is to use a PLL as the demodulator, with carefully chosen loop bandwidth and damping. Usually, the loop bandwidth is less than the IF bandwidth, and this reduces the instantaneous system noise bandwidth, which improves CNR. (Such PLLs also tend to reject quadrature noise.) In practice, this can be used to lower the FM threshold several dB, as the Plessey SL1451 demod does (8).

Another way to achieve threshold extension is to employ a narrower IF filter than would normally be used with this FM TV format, and use the demodulated FM video information to deviate the receiver's local oscillator slightly in the opposite sense. The FM signal passing through the IF filter then has had its deviation reduced (by the amount of FM introduced to the LO). This narrower filter has less noise bandwidth, and the CNR at the demodulator is again improved enough to allow operation with a somewhat smaller antenna (8).

Obviously, there are other considerations — you are putting feedback

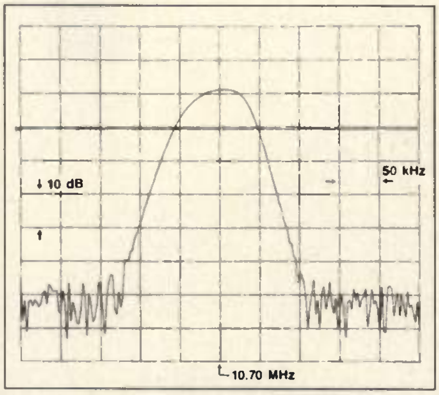

Figure 4b. Amplitude vs. frequency for same cascaded filters.

around the time delay of the IF filter and servo design rules apply, but that is the basic idea. Again, a somewhat narrower passband "chases" the FM carrier, instantaneously reducing its peak deviation and bandwidth. As with PLL demodulation, the final selection of circuit constants and bandwidth tends to be somewhat empirical.

Application of Feedback to FM Broadcast Demodulation

A similar idea was tried to selectively reduce the deviation of high quality, interference free FM stereo signals to the point where a single, very narrow, high selectivity IF filter could be used. Demodulated audio from the CA3089 FM detector was fed back to a very linear 98.7 to 118.7 MHz local oscillator such that peak deviation could be reduced on demand from the normal 75 kHz, to about 50 kHz at the mixer output. See Figure 1.

When narrower bandwidth is needed, the feedback is reduced or turned off. This is how the IF bandwidth is modified as reception conditions change. The advantage over several discrete IF filters is that only one is needed, and any degree of bandwidth between two extremes is available. Noisy reception may then be cleared up by adjusting the tuning and bandwidth controls.

Design Details

An important block is the local oscillator of this superhet receiver. To produce a 10.7 MHz IF when receiving the FM band from 88 to 108 MHz, it has to tune from 98.7 to 118.8 MHz (so called high side injection). More importantly, it is the summing junction where the demodulated audio is applied to the incoming signal which reduces its deviation. Any voltage to frequency non-linearity will be

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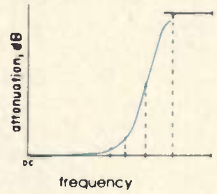
LOW PASS



low pass, Plug-in, dc to 1200MHz

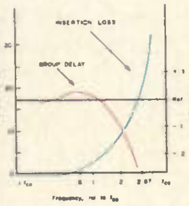
Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
★LP-5	DC-5	8-10	10-200
★LP-10.7	DC-11	19-24	24-200
★LP-21.4	DC-22	32-41	41-200
★LP-30	DC-32	47-61	61-200
★LP-50	DC-48	70-90	90-200
★LP-70	DC-60	90-117	117-300
★P-90	DC-81	121-137	167-400
★LP-100	DC-98	146-189	189-400
★LP-150	DC-140	210-300	300-600
★LP-200	DC-190	290-390	390-800
★LP-250	DC-225	320-400	400-1200
★LP-300	DC-270	410-550	550-1200
★LP-450	DC-400	580-750	750-1800
★LP-550	DC-520	750-920	920-2000
★LP-600	DC-680	840-1120	1120-2000
★LP-750	DC-700	1000-1300	1300-2000
★LP-800	DC-720	1080-1400	1400-2000
★LP-850	DC-760	1100-1400	1400-2000
★LP-1000	DC-900	1340-1750	1750-2000
★LP-1200	DC-1000	1620-2100	2100-2500

Price, (1-9 qty), all models: plug-in $14.95, BNC $32.95, SMA $34.95, Type N $35.95

Surface-mount, dc to 570MHz

Model No.	Passband MHz loss < 1dB	Stopband, MHz loss > 20dB	Stopband, MHz loss > 40dB
SCLF-21.4	DC-22	32-41	41-200
SCLF-30	DC-30	47-61	61-200
SCLF-45	DC-45	70-90	90-200
SCLF-135	DC-135	210-300	300-600
SCLF-190	DC-190	290-390	390-800
SCLF-380	DC-380	580-750	750-1800
SCLF-420	DC-420	750-920	920-2000

Price, (1-9 qty), all models: $11.45

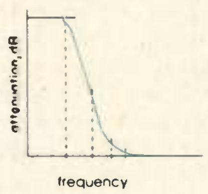


Flat Time Delay, dc to 1870MHz

Model No.	Passband MHz loss < 1.2dB	Stopband MHz loss > 10dB	Stopband MHz loss > 20dB	VSWR Freq. Range, DC thru 0.2fco X̄	VSWR Freq. Range, DC thru 0.6fco X̄	Group Delay Variations, ns Freq. Range, DC thru fco X̄	Group Delay Variations, ns 2fco X̄	Group Delay Variations, ns 2.67fco X̄
★BLP-39	DC-23	78-117	117	1.3:1	2.3:1	0.7	4.0	5.0
★BLP-117	DC-65	234-312	312	1.3:1	2.4:1	0.35	1.4	1.9
★BLP-156	DC-94	312-416	416	0.3:1	1.1:1	0.3	1.1	1.5
★BLP-200	DC-120	400-534	534	1.6:1	1.9:1	0.4	1.3	1.6
★BLP-300	DC-180	600-801	801	1.25:1	2.2:1	0.2	0.6	0.8
★BLP-467	DC-280	934-1246	1246	1.25:1	2.2:1	0.15	0.4	0.55
▲BLP-933	DC-560	1866-2490	2490	1.3:1	2.2:1	0.09	0.2	0.28
▲BLP-1870	DC-850	3740-6000	5000	1.45:1	2.9:1	0.05	0.1	0.15

Price, (1-9 qty), all models: plug-in $19.95, BNC $36.95, SMA $38.95, Type N $39.95
NOTE: ▲: -933 and -1870 only with connectors, at additional $2 above other connector models.

HIGH PASS

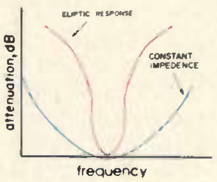


high pass, Plug-in, 27.5 to 2200MHz

Model No.	Stopband MHz loss < 40dB	Stopband MHz loss < 20dB	Passband, MHz loss < 1dB	VSWR Passband Typ.
★HP-25	DC-13	13-19	27.5-200	1.8:1
★HP-50	DC-20	20-26	41-200	1.5:1
★HP-100	DC-40	40-55	90-400	1.8:1
★HP-150	DC-70	70-95	133-600	1.8:1
★HP-175	DC-70	70-105	160-800	1.5:1
★HP-200	DC-90	90-116	185-800	1.6:1
★HP-250	DC-100	100-150	225-1200	1.3:1
★HP-300	DC-145	145-170	290-1200	1.7:1
★HP-400	DC-210	210-290	395-1600	1.7:1
★HP-500	DC-280	280-365	500-1600	1.8:1
★HP-600	DC-350	350-440	600-1600	2.0:1
★HP-700	DC-400	400-520	700-1800	1.6:1
★HP-800	DC-445	445-570	780-2000	2.1:1
★HP-900	DC-520	520-660	910-2100	1.8:1
★HP-1000	DC-550	550-720	1000-2200	1.9:1

Price, (1-9 qty), all models: plug-in $14.95, BNC $36.95, SMA $38.95, Type N $39.95

BANDPASS



bandpass, Elliptic Response, 10.7 to 70MHz

Model No.	Center Freq. (MHz)	Passband I.L. 1.5 dB Max (MHz)	3 dB Bandwidth Typ. (MHz)	Stopbands I.L. > 20dB at MHz	Stopbands I.L. > 35dB at MHz
★BP-10.7	10.7	9.6-11.5	8.9-12.7	7.5 & 15	0.6 & 50-1000
★BP-21.4	21.4	19.2-23.6	17.9-25.3	15.5 & 29	3.0 & 80-1000
★BP-30	30.0	27.0-33.0	25-35	22 & 40	3.2 & 99-1000
★BP-60	60.0	55.0-67.0	49.5-70.5	44 & 79	4.6 & 190-1000
★BP-70	70.0	63.0-77.0	68.0-82.0	51 & 94	6.0 & 193-1000

Price, (1-9 qty), all models: plug-in $18.95,
BNC $40.95, SMA $42.95, Type N $43.95

Constant Impedance, 21.4 to 70MHz

Model No.	Center Freq. MHz	Passband MHz loss < 1dB	Stopband loss > 20dB at MHz	VSWR 1.3:1 Total Band MHz
★IF-21.4	21.4	18-25	1.3 & 150	DC-220
★IF-30	30	25-35	1.9 & 210	DC-330
★IF-40	42	35-49	2.6 & 300	DC-400
★IF-50	50	41-58	3.1 & 350	DC-440
★IF-60	60	50-70	3.8 & 400	DC-500
★IF-70	70	58-82	4.4 & 490	DC-550

Price, (1-9 qty), all models: plug-in $14.95,
BNC $36.95, SMA $38.95, Type N $39.95

NOTE: **★Add Prefix P, B, N, or S for Pin, BNC, N, or SMA connector requirement.**

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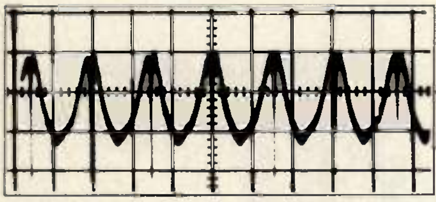

Figure 5a. Demodulated output without feedback.

transferred to the demodulated output and increase audio output distortion.

A highly linear VCO was designed based on M/A-COM's design data and nomograph in their catalog SP-101 "Semiconductor Products Master Catalog" (9).

The LC tank for the oscillator has a resonant point proportional to $1/\sqrt{LC}$. If the L is a fixed inductor, then C (usually a varactor diode used to tune the circuit electronically) must exhibit a voltage squared capacitance law for linear tuning in MHz per volt. Special doping profiles within the diode are used to produce given values for Gamma, the capacitance voltage slope exponent.

When this number, Gamma, is a constant over the tuning range, it is possible to design a linear VCO. When equal to two, the voltage-squared-law mentioned above is obeyed exactly. For other values of constant Gamma, M/A-COM has a nomograph (on page 5-29 of the catalog mentioned previously) that was used to produce the following design.

1. Ratio of Fmax/Fmin — 98.7 to 118.7 MHz in our case.
2. Diode Gamma required — 1.5 was chosen from the nomograph.
3. Necessary diode capacitance ratio Cmax/Cmin — 2:1, which was chosen as 4.5 to 9.0 pF for the particular diode we used, a GaAs hyperabrupt-junction part.
4. Circuit coupling factor — A parameter determined by an additional series or shunt fixed capacitance that determines the degree of control the varactor has over the resonant frequency. In this case, 6 pF shunt C was taken from M/A-COM design curves.

A tuning inductance of 178 nH was needed with the above values; we used a 180 nH fixed coil.

The oscillator circuit, as implemented, is shown in Figure 2. The 6 pF circuit coupling capacitor was made from two 10 pF NPO caps in series, allowing 1 pF for stray C. Proper phase for oscillation with an inverting active device was obtained by grounding the center tap of the tank as shown. The MiniCircuits MAE8 is an ideal active device for oscillator design because it is self biasing, has a low output Z, and a very high input Z that does not load down the resonant circuit, as a matched device would. This linear VCO design method produced a prototype that met all expectations at first turn on, and no changes were required to part values or the schematic.

The circuit tunes linearly over the required band from 4.5 to 9.0 volts. The slope is such that 0.004 V audio feedback in the proper phase reduces 75 kHz incoming FM deviation to 50 kHz, which is the design goal. (See the V-F curves in Figure 3). A resistive attenuator network sums the audio with the channel select tuning voltage. Nonlinearity was low enough that it did not to compromise audio distortion to any great degree. However, low distortion audio grade test equipment was not available at the time to confirm how low.

IF Filter and Stability Considerations

The most difficult part of this design was deciding whether the final design would be stable when feedback was applied around the system. To do a Bode plot requires accurate models of each circuit element: little group delay data was published on the parts I intended to use, and a highly detailed paper analysis was not attempted. One SAW IF filter candidate had 21 uS average delay across its passband. Clearly, at about 23.8 kHz, this represents 180 degrees phase shift. If there were another time delay and/or phase shift in the system, (and there certainly is!), this would reverse the sense of the negative feedback we are trying to apply and cause

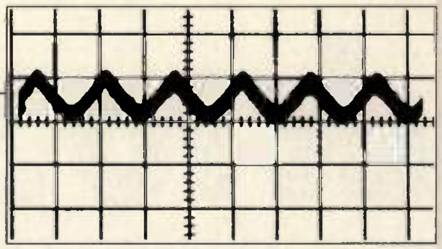

Figure 5b. Demodulated output with feedback.

gain peaking, or in the extreme, oscillation.

The best compromise of IF filter delay vs. narrow bandwidth I found resulted from the cascade of two Murata group-delay-compensated ceramic filters of about 150 kHz BW each. Each filter, part number SFE10.7Z, had 3.5 uS measured average delay. The two filter cascade provided a close-in rapid rate of rolloff and ultimate rejection of almost 70 dB total (see Figure 4). Without feedback, this filter is too narrow for broadcast FM reception. A ground plane is mandatory to achieve this level of ultimate rejection.

An eight µS delay in the IF strip alone corresponds to 180 degrees phase shift at 62 kHz. More delay anywhere in the system will reduce the phase crossover point toward the FM stereo baseband limit of 53 kHz. At this time no attempt was made to modify the gain margin via lead compensation. Typically, we need very little feedback to reduce 75 kHz deviation to 50 kHz. At this time it was decided to just try this concept rather than design in phase compensation based on incomplete models.

It is important to note that this is really a form of frequency-locked loop or Automatic Frequency Control system. Frequency is the controlled and controlling variable. In contrast, a phase locked loop integrates system phase to produce a frequency control effect. Such an integrator adds another 90 degrees of phase shift and must be accounted for in stability calculations. It really is excusable, on a first cut, to simplify the analysis we used in the receiver design above since neither an integrator nor phase detector is included.

Receiver Design

The IF strip consists of the cascaded filters with MRF901 buffer amplifiers. They feed a CA3089 FM demodulator. The voltage controlled oscillator with its audio input and a 10 turn channel selection (tuning) pot is used to generate the required LO frequency. The mixer is a MiniCircuits SRA 1 dBm. Avantek MSA 1105 MMIC amplifiers are used in two places: to isolate and amplify the LO to the required power level, and to terminate the mixer output in a resistive 50 ohm load. This is necessary to terminate possible birdies (spurious responses)

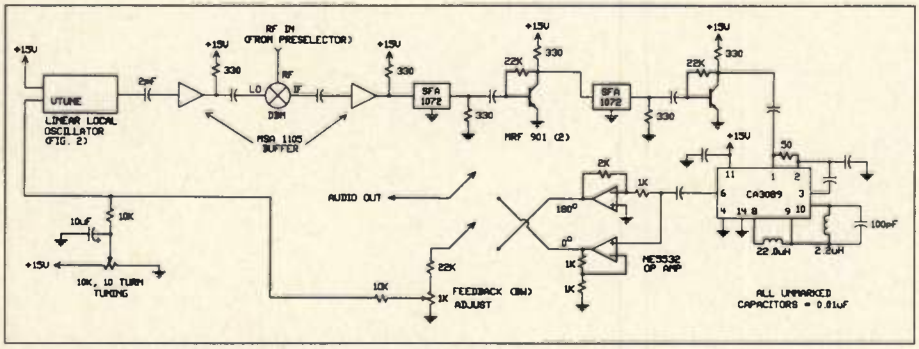

Figure 6. Schematic diagram of the experimental receiver.

caused by unwanted mixer products reflecting back into the mixer's output, a problem well known to RF designers.

A simple preselector (not shown here) was used to gain a degree of image rejection from the other mixer sideband at +21.4 MHz (twice the IF of 10.7 MHz in this case) when off the air reception was attempted. The demodulated output was raised to earphone level with an 5532-type op amp that produced both audio polarities. This was necessary since it was not known what absolute polarity the 3089 audio output had. The normal 75 µS deemphasis network, if used, has to be placed after the feedback point as not to degrade phase margin within the loop any further. A schematic of the receiver circuit is shown in Figure 6.

Bench Test

A 100 MHz signal generator was FM'ed ±75 kHz with a 15 kHz triangle wave and was applied to the receiver with the feedback off. As expected, the waveform was distorted at the deviation limits. Advancing the feedback control reduced the amplitude such that the clipping was stopped (see Figures 5a and 5b). A frequency response plot was then run from 50 Hz to 100 kHz.

The point of maximum peaking was 4 dB at 49 kHz using the feedback setting from above. This indicates that flat response from 50 Hz to 53 kHz (for stereo reception) will require that the feedback system be lead compensated. The Plessey feedback demodulator mentioned above (SL1455) requires such compensation. The required amount depends on whether PAL-4.43 color video or NTSC is received, since the bandwidths are different. Since low distortion audio grade test equipment was not available, the next part of the testing was via ear.

Air Test

A Finco FM-9G antenna and mast-mounted preamplifier was used as the signal source. The antenna was installed about 25 feet above average local terrain and was oriented to the northeast, towards New York City, approximately 40 miles distant. A 12 year-old Kenwood KT 7500 dual bandwidth tuner was used as a control. Using the 7500, most stations were received cleanly using only the receiver's narrow filter. In nine cases, the bandwidth could be opened up by switching to the wide position. The feedback receiver produced similar results, but at varying feedback (bandwidth) settings, indicating the continuous bandwidth idea allowed some optimization of the passband for the best possible reception. This is a subjective method and can not easily be duplicated via precision instrument techniques.

Too wide a setting (too much feedback) causes complete breakup with a sudden onset.

Conclusions

An old technique, feedback FM demodulation, was found to work as an alternative method of providing user-selectable IF strip bandwidth, which can optimize FM stereo reception. Development is by no means complete; the feedback parameters need optimization, stereo reception was not attempted, nor was extensive objective testing done, but at this time the utility of the technique seems clear. A simple feedback arrangement was shown to affect a single-knob IF bandwidth control in this home lab experiment.

It is not clear whether the loop dynamics actually degrade strong-signal noise floor (deviation is being reduced, and FM systems depend on deviation squared to affect their "FM improvement") (10). I plan to continue to experiment with this arrangement as time permits. *RF*

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About the Author

Jerry Pulice received his BS in Engineering Science from Richmond College, CUNY and his MSEE from Fairleigh Dickenson University. He has been an RF/analog engineer with Smith's Industries since 1988. His work has included fiber optic and copper distribution of video and GPS RF signals, avionics, multi-loop synthesizers, CATV security devices and a multistandard video game color encoder. He can be reached at SIFPD, 7-11 Vreeland Dr., Florham Pk., NJ 07932, or by phone at (201) 822-1300 x2689.

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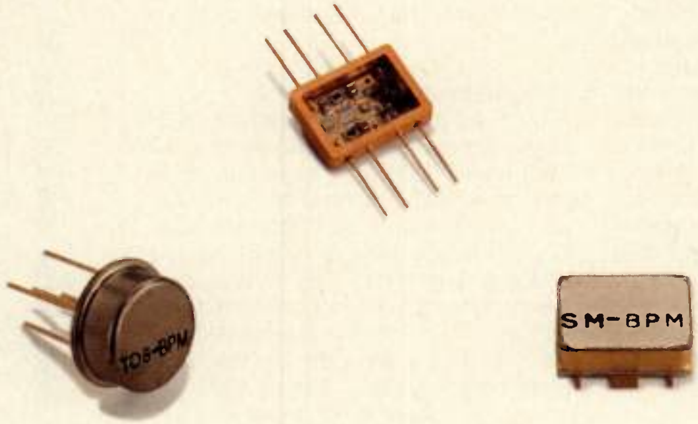

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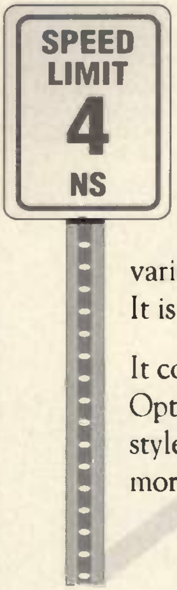

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High Dynamic Range Tunable Bandpass Filters for Communications

By Richard Markell
Linear Technology Corp.

Octave or decade wide bandpass filters are non-trivial to design. This is because the bandpass output of the state-variable filter configuration, be it switched capacitor or active RC, severely limits the achievable dynamic range. Thus, wide bandpass filtering requires the use of a highpass filter at the input in series with a lowpass filter to achieve the desired specifications.

Wideband bandpass filters occupy a niche in the communications arena of signal processing. The wideband bandpass function is required in receiver IF applications which traditionally used the crystal filter and/or active RC's. Sonar applications also demand steep, wide, low noise bandpass filters to allow analysis of "chunks" of the frequency spectrum, one at a time or, perhaps more likely, in parallel.

Recent improvements in switched capacitor filter technology allow designers the luxury of using switched capacitor filters in these applications. True clock tuning allows variable bandwidth filters to be implemented with only a few parts.

This article details the design of a wideband (10 kHz-100 kHz) bandpass filter using a single LTC1064 plus an LTC1064-4. The filter may be tuned ratiometrically by simply changing the clock frequency. As an example, by simply changing the clock frequency, the filter may be operated from 1 to 10 kHz or 100 Hz to 1 kHz, and so on.

The LTC1064 is a quad universal switched capacitor filter while the LTC1064-4 is the same basic chip but with integrated resistors configured to implement an 8th order elliptic low-pass filter. Both filters have low noise and can operate to 100 kHz. The combination bandpass filter has a set of tough design specifications: total integrated noise in the passband less than 200 microvolts, passband ripple less than .4 db or ± .2 db, and steep rolloffs at the band edges (–70 db at 5 kHz, and –70 db at 200 kHz).

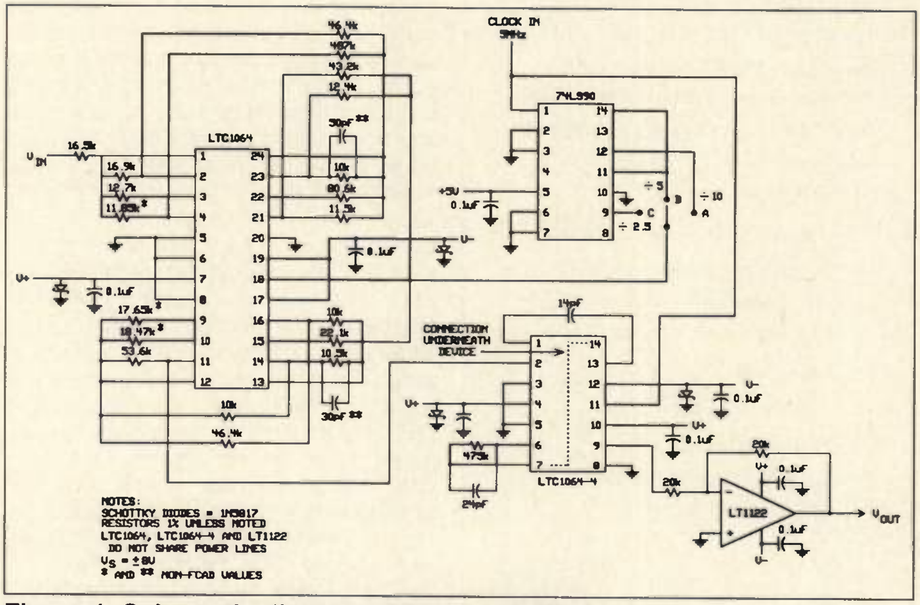

Figure 1. Schematic diagram of bandpass filter.

Filter Trade-Offs

RF designers generally design filters using inductors and capacitors or, if the filter is at a lower frequency, using op amps, resistors and capacitors. The limiting factor in these types of designs is

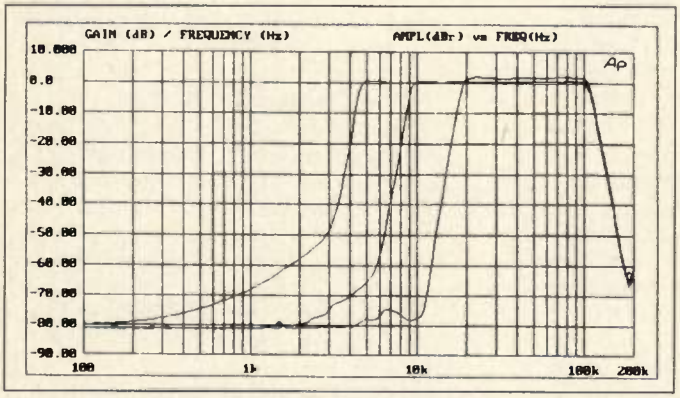

Figure 2. LTC 1064/LTC 1064-4 bandpass response with high pass filter cutoff frequency varied. V_{in} = 2.2 V rms. V_s = ±7.5 V.

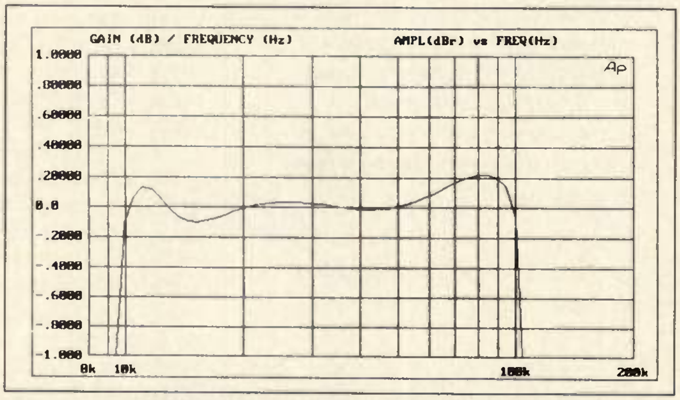

Figure 3. Passband detail, LTC 1064/LTC 1064-4. 10 kHz - 100 kHz BPF, V_s = ±7.5 V, V_{in} = 2.2 V rms.

the sensitivity and accuracy of the capacitors and inductors. A high Q filter section requires precise (generally .1 percent or better) components which are either too expensive or impossible to obtain. Additionally, the physical printed circuit board area required for a complex filter implemented with op amps or passive components is considerable.

The switched capacitor filter is not a newcomer to the filter party. It has been used in telephony systems for years, albeit in non-specification intensive situations. Recent advances in switched capacitor filter technology have greatly improved their specifications, allowing single 24 pin DIP or 16 pin SO (surface mount) packages to replace multiple op-amp circuitry in all but the most noise critical applications. As a tangible example, a switched capacitor 8th order Butterworth filter is available as the LTC1064-2. This filter provides clock tunable corner frequencies to 140 kHz, 80 uV RMS total wideband noise, stopband attenuation better than 74 dB (at 3 fc) and a typical THD (Total Harmonic Distortion) specification of 0.03 percent or better.

In addition, software is now available to optimize switched capacitor filter design for lowest THD or lowest noise performance. This software (LTC's FilterCAD) frees the designer from the mathematical exercises which often are not dear to the heart of the busy system engineer. The software program not only can be very useful for filter design and calculation of the resistors required for implementation, but it also contains algorithms to ensure optimization of the dynamics of the switched capacitor filter devices so that no node of the part can saturate before the desired output node.

Design

The topology for the bandpass filter combines an LTC 1064-4 elliptic lowpass filter with an LTC1064 elliptic highpass. The highpass filter was designed using Linear Technology's FilterCAD filter design software. The LTC1064-4 gives attenuations of greater than 70 db at 2 times cutoff; the highpass filter was designed to be almost the mirror image of the lowpass filter. Figure 1 shows the schematic of the composite bandpass filter. The 74LS90 is used as a divide-by-5 to clock the LTC1064 10 kHz highpass filter. In addition to providing both 5 MHz and 1 MHz clocks, this circuit allows the clocks to be synchronous. This is essential for well behaved operation of sampled data filters.

Test Results

Figure 2 shows the overall frequency response of the bandpass filter at highpass corners of 5, 10 and 20 kHz. Note the steep slopes at the transition regions of the filter. The measured noise in the passband of the filter was 160 microvolts RMS. This translates to a dynamic range (with a 2.2 volt RMS input) greater than 80 dB. Passband ripple (Figure 3) shows excellent specifications of less than ± .3 db. It should be noted that the values of the resistors marked with an asterisk in Figure 1 were modified slightly from the values given by FilterCAD. The passband ripple using the FilterCAD values measured better than ± .4 db. An additional note should be made of the 50 picofarad and 30 picofarad capacitors shown with double asterisks. These capacitors serve as RC filters on two of the highpass outputs of the LTC1064 to roll off the response of the HPF (well past the passband of the overall filter) to limit noise aliasing back to the filter's passband. These capacitors may limit the overall tunability of the filter or they may need a range changing switch.

Conclusions

A 16th order low noise, high quality, bandpass filter can be designed with only 4 active components and a handful of resistors. The filter meets specifications that can only be approached with the most sophisticated hybrid filter solutions. Further, the filter was primarily designed with the FilterCAD program, so that the designer may change parameters with the push of the "Enter" key. This bandpass filter only touches on sophisticated designs, formerly the domain of wise mathematicians. These filters can be designed and implemented with Linear Technology's switched-capacitor filters and the FilterCAD software package. ***RF***

About the Author

Richard Markell is Applications Manager at Linear Technology Corporation where he has been working for 4-1/2 years. At LTC he has specialized in the design of signal processing systems for customer applications. These designs have included all types of switched capacitor filter designs, A/D system design and power supply design. He may be reached at Linear Technology Corporation, 1630 McCarthy Blvd., Milpitas, CA 95035. Tel: (408) 954-8400.

Noise Generators Find Key Uses in VXI Based Systems

By Edward J. Garcia
Noise Com, Inc.

Noise sources have proven to be a valuable component in ATE applications. Their programmable output levels and bandwidths have been integrated into IEEE-488 bus controlled systems with great effectiveness. Now, VXIbus controlled systems are finding growth in the RF and microwave frequency ranges, which has facilitated the need for a new line of VXIbus controlled noise sources. Currently, there are already hundreds of VXIbus products available with many manufacturers producing RF and microwave products.

The VXIbus is a complete modular standard which offers numerous advantages over other bus controlled standards. It is designed to be independent of manufacturer and system hierarchy, therefore the user is not bound to a specific operating system or a particular microprocessor.

Standardization is required for effective downsizing of both commercial and military test equipment. The VXIbus, unlike its predecessor, the IEEE-488, is not just an interface standard, but a modular standard which incorporates common power supplies, cooling systems and a common chassis. This concept of shared resources shrinks component size dramatically and enables modules to be coupled close together eliminating long cable connections and their associated propagation delays and timing uncertainties. The result is a downsized test instrument or test station with several orders of magnitude of increased interface speed enabling higher level components to be integrated together. The elimination of redundant support components improves the reliability of a system. In addition, the modularity increases serviceability. It also provides good re-configuration capabilities, since instrument modules are easily removed and installed. Unique capabilities can be incorporated in a VXIbus based system without designing specialized hardware. This is achieved by linking several instruments together to perform a specialized function. This is known as a virtual instrument. The user interfaces with the instrument cluster as if it were a single unit. The virtual instrument contains a command instrument, which controls one or more servant instruments. The interaction between the command instrument and the servant is transparent to the end user.

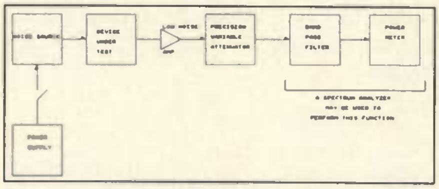

Figure 1. Noise figure measurement using the substitution method.

Early applications of VXI were mostly military, but the portability and flexibility of VXI is becoming more attractive to commercial users as well. A VXI test station can be easily transported for on-site testing and a VXI chassis can be configured to operate on vehicle power if mobile site testing is required.

Since many VXIbus systems interface with an IEEE-488 bus controller, they can be limited in their speed due to this architectural choice. Regardless, increased throughput results due to the memory mapped architecture of VXIbus. This direct access to the registers is unencumbered by intervening protocol and can result in at least a factor of 10 increase in throughput.

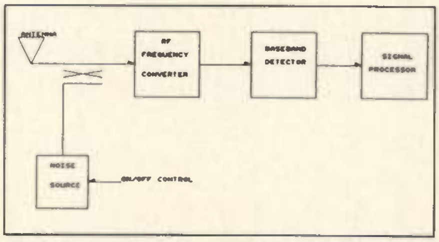

Figure 2. Using a noise source for receiver bite.

Until recently there was no easy way to integrate a noise source into a VXIbus system. Since there is now a standard line available for users, it will be helpful to discuss the fundamentals of noise sources and their applications.

Noise Applications

Noise sources are specified in terms of their output power and frequency content. For calibrated low level noise sources, output is expressed in terms of Excess Noise Ratio (ENR):

$$ENR = 10 \bullet LOG[(Th - 290)/290] \qquad (1)$$

with Th being the equivalent noise temperature (in Kelvins, K) of a matched resistor. Qualitatively, ENR is simply a measure of how much the output of the noise source is above the thermal noise floor. From Equation 1, one can see that an ENR value may be negative for output levels near thermal noise. For higher output levels, ENR is approximately the number of decibels above −174 dBm/Hz that a noise source produces.

Noise sources are commonly used to inject a calibrated amount of noise into a system or a device and measure its response. The noise figure and gain are two key parameters which can be measured and monitored using noise sources. Figure 1 depicts a noise source being used to measure the noise figure of an amplifier. There are several techniques for measuring noise figure, but one of the most accurate is the substitution method. This method eliminates the non-linearity uncertainty of the power measuring device. Accuracy is dependent only on the variable attenuator. A reference point is set on the power detector with the noise source de-activated. The noise source is then turned on and the variable attenuator is increased until the power reading is the same as when the noise source was off. The measuring device can be a spectrum analyzer or frequency selective power

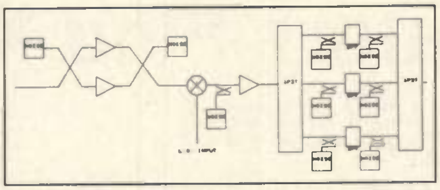

Figure 3. Fault isolation using noise sources.

meter. The change of the attenuator setting is the Y factor. The noise figure of the amplifier is then calculated as follows:

$$Y = 10^{(Y(dB)/10)} \quad (2)$$

$$NF(dB) = ENR(dB) - 10 \cdot LOG(Y-1) \quad (3)$$

Where ENR is the excess noise ratio of the calibrated noise source. If the contribution to noise figure by the low noise amplifier following the device under test is significant, then this second stage effect should be subtracted from the above calculation to arrive at the noise figure of the amplifier.

Low output noise sources are being built into digital and analog receivers to provide continuous on-board monitoring of various parameters as shown in Figure 2. A typical noise source designed for Built-In-Test (BIT) has an ENR of 15 to 30 dB. This device can be switched into the signal path at the receiver input to monitor receiver integrity. A coupler can also be used to inject the noise, in which case a higher output noise source may be used to compensate for the coupling loss. Since the noise source output level is known, it can be turned on and the gain of the receiver calculated. By switching the noise source on and off, the noise figure of the receiver can be calculated from the difference in the measured output power, again using the Y-factor as calculated in equation 2 of the above example:

$$\text{Noise Figure} = 10 \cdot LOG[(Y-1)/(Y+1)] \quad (4)$$

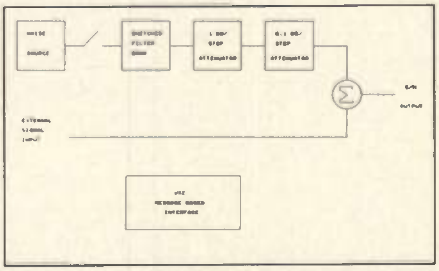

Figure 5. VXIbus based noise instrument.

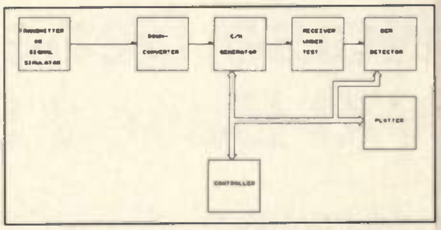

Figure 4. Bit error rate test setup.

Since noise sources are very broadband, the frequency response of a system can also be monitored for gain flatness.

Extending the concept of built-in noise sources provides fault isolation capability. Figure 3 depicts a portion of a receiver with a noise source integrated with each component. By selectively turning on the noise sources, the system controller can effectively isolate degraded or failed components. System performance can also be analyzed by using the measured gain values associated with each noise source. The noise source can be hybridized into the receiver component, thereby not consuming any additional space in the system. The small size and low cost of the noise sources make this a very cost effective method of fault isolation. Noise sources with these output levels are available in chip form and in packages ranging from surface mount devices, TO-8 cans and dual in-line packages to coaxial structures, all of which can be easily integrated into VXIbus based receivers and subsystems.

The new VXIbus based noise instruments are higher power noise sources which are specified in terms of overall output power instead of ENR. They are basically low level noise sources amplified to levels which can be filtered and attenuated based on the application. Since noise has a constant power spectral density over frequency, the power measured will be a function of the measurement device bandwidth. When this noise signal is injected into a device or system, the power into the device is reduced if its bandwidth is smaller than that of the noise source. For example, if the noise source bandwidth is 100 MHz and its output is +10 dBm, then its spectral density is –70 dBm/Hz. From this value the power into any bandwidth can be calculated. If the same noise source is injected into a device with a bandwidth of 10 MHz, the effective injected power becomes 0 dBm since the bandwidth has been reduced by a factor of 10. The spectral density, however, remains at –70 dBm/Hz. These amplified noise sources are extremely valuable when summed with an input signal because they can be used to create variable carrier to noise ratios (C/N) which are required to test radios. The performance of digital radios is specified in terms of its Bit Error Rate (BER) under various C/N conditions. The signal to noise ratio is typically expressed in several forms, including bit energy to noise density (E_b/N_o), and carrier power to noise density (C/N_o) in addition to C/N. These ratios can be negative during normal operation, for example with spread spectrum systems where the carrier is modulated by a pseudo-random code at a high clock rate. Figure 4 depicts a setup for measuring Bit Error Rate. A typical measurement will vary the ratio at or around the receiver threshold and yield a plot of the system BER versus the signal to noise ratio.

Applications of VXIbus Controlled Noise Instruments

The new line of available VXIbus based noise instruments will aid in many applications including BER testing, multipath fading simulation, component analysis, radar test beds, as well as self-test and calibration of the VXI system. Figure 5 depicts a VXI noise instrument. The module consists of a high level noise source followed by a switched filter bank, step attenuators, and a combiner for the user injected signal. This module provides good flexibility with programmable output power from +13 dBm to the thermal noise floor and user specified bandwidths. It utilizes a message based VXIbus interface, and is compatible with the recently released Revision 1.4 of the VXIbus System Specification. The internal components are housed in RFI tight enclosures with power line and control line filtering. The noise output can be used to calibrate and functionally test other devices in a VXI system, eliminating the need to return modules to a calibration facility, and therefore eliminating system down time. Types of self-tests which may be performed with the noise instrument include calibration of power measuring devices, amplitude response of components over frequency and time, and receiver characteristics such as sensitivity, saturation, intermodulation distortion, gain, noise figure, and detector response.

By integrating a power meter and a noise source into a virtual instrument, a complete precision carrier to noise generator is realized. The power meter is used to set the C/N ratio by measuring the noise power and the carrier power. The noise density can be computed from the power measurement and the noise bandwidth of the noise instrument. Computational corrections for bandwidth, data rates, etc. can be performed transparently to the user. The substitution method described above can be utilized if the accuracy of the power meter is not sufficient for the desired C/N accuracy. The switched filter bank residing in the noise instrument is user specified to allow testing in different frequency bands, or may be bypassed for unfiltered broadband applications. The noise bandwidth of the filters can be precisely calibrated such that noise density can be derived from the measured noise power.

Conclusion

VXIbus based hardware is gradually becoming accepted in the industry as users become more familiar with the devices which are available and, more importantly, gain an understanding of the advantages of VXI systems. Manufacturers are also beginning to see the advantages and as a result, many new instruments are becoming available. One line of new VXIbus based instruments are noise sources which are very flexible and can be utilized for calibration, self test and fault isolation, built-in testing, as well as higher level system testing such as BER and receiver characterization. The continued acceptance of this relatively new instrumentation platform will bring even more incentives for using VXI. ***RF***

About the Author

Edward J. Garcia graduated from Lehigh University with a BSEE. He worked as a microwave engineer at KDI/Triangle prior to joining Noise Com in 1990. He currently is the Chief Engineer, overseeing production engineering and new product development. He can be reached at Noise Com, Inc., Paramus, New Jersey 07652. Tel: (201) 261-8797

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EEsof, in partnership with Sawtek has included Sawtek's 70 MHz and 160 MHz standard, off-the-shelf SAW filter families in EEsof's Series IV Vendor Component Libraries. The 140 MHz standard filter family will follow.
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RF literature

SAW Catalog
Sawtek has released its 1993 Product Catalog, its first catalog ever. In addition to standard product listings, the catalog includes how to specify a custom SAW filter, a section on application-specific SAW devices, and PC board layout tips to assist the RF designer.
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Trimmer Capacitors
Voltronics has published an eight-page catalog describing its new line of high performance, half-turn ceramic trimmer capacitors. There are nine different styles varying in size from 2.4 mm to 8.5 mm in diameter. Tuning ranges are from 2 to 40 pF maximum for either printed circuit or surface mount varieties.
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RF product report

Signal and Sweep Generators Re-Aimed at Commercial Market

By Andy Kellett
Technical Editor

Signal and sweep generators are essential to lab testing of RF components and systems. Occupying a role analogous to a wind tunnel in aeronautical engineering, signal and sweep generators provide controlled, electronic tests of RF circuits and systems. Because test signals are such an essential part of RF testing, these instruments are used all across the RF field.

The basic tests performed with signal and sweep generators are the same as they have always been: receiver testing, LO substitution, and device and system characterization, among others. What has changed is the market in which these tasks are done.

RF manufacturers are selling less to military customers and trying to sell more to the growing number of commercial customers. With the shift in market comes new considerations. "Companies are having to learn how to do a whole new business where there are a lot of new driving factors, particularly cost. Given that, we are all babes in the woods again, starting out in a new market," says Donn Mulder, Product Marketing Manager at Anritsu Wiltron Sales Co.

What to sell in a new market is one of the first decisions to be made. Wiltron's Mulder notes that customers want to buy signal generators and sweepers that do what they need them to do, no more, no less. Special capabilities can be offered as options. This is what Marconi Instruments has done with their 2030/2040 line of signal generators. "We offer different options, each of which represents a vertical market," says Marconi President Carl Pepple. For instance, Option 6 for the Marconi 2030 adds capabilities for testing avionics.

Special features aside, users are most interested in frequency range and stability, power output, harmonics, phase noise and modulation capabilities according to Dave White, Vice President of Marketing and Sales at Gigatronics. The frequencies in which tests are done reflects where in the spectrum most of the activity lies. Part 15 applications are stimulating generator activity at 2.4 GHz.

Frequency stability, harmonics and phase noise are functions of the method used to generate the carrier frequency. Fractional-N synthesis is the predominant carrier generation method into the UHF range. Many of the companies that make generators that reach above the UHF band use YIG oscillators, mixing them down to reach the HF and VHF ranges and multiplying them to get higher. Very accurate, very low noise signal sources are made by Programmed Test Sources by using very stable SC-cut crystals and direct synthesis. Signal generator features such as modulation capabilities can be added as options to these sources.

Direct digital synthesis (DDS) is known for its frequency agility and waveform flexibility. DDS generators from Stanford Research, Analogic and others are still in the tens of MHz, but they find use in LO applications and can be mixed up to higher RF frequencies. "Everyone is worried that with ones and zeros running around that there will be glitches everywhere," says John Willison, Vice President of Engineering at Stanford Research. "but our spurs are –70 dB down."

Signal and sweep generator control is another specification that customers and manufacturers focus on. "Button-per function" control is the dominant control panel architecture. "The instrument before our new 2200/2210 sweeper was menu driven. The reason that was done was to bring price down; it eliminates a lot of hardware. The customer base just didn't like it, so we went back to the button per function design," says Boonton Electronics Sales Manager, Richard Anlas. The readability and ruggedness of front panels are also important when considering generators that will be used in the field. "We're using low current LED displays so that you can use them in any light conditions, and we use tactile front panels. They have two distinct advantages, they are have very, very low radiated emissions and the other is reliability," says Ken Harrison, National Sales Manager for Wayne Kerr.

More signal and sweep generators are being controlled by computer. Automatic test equipment is slowly working its way onto manufacturing floors. Nearly all signal and sweep generators possess an RS-232 or GPIB bus as at least an option. These instruments generally retain the front panel and front panel controllability. Eliminating the front panel reduces the space an instrument occupies, and size reduction is one of the goals of the VXIbus standard. VXIbus instruments are exclusively computer controlled and contained in racks much smaller than the traditional 19 inch, rack-and-stack types.

Like the much of the rest of the RF test and measurement industry, signal and sweep generator manufacturers have seen less business over the past few years. Military spending declined while a recession made companies more sluggish and less able to adapt to changing conditions. For whatever reason the market slowed down, it seems to be recovering now. According to Russ Byrd, Wavetek Product Marketing Manager, business at his company has been steady, and they are beginning to sell replacements to old equipment. Gigatronics' Vice President of Marketing and Sales, Dave White, says sales of their 7000 series of multiband signal generators is growing, and that sales overall have been strong.

There will be no shortage of potential customers for signal and sweep generators in the future. Whether those potential customers become real buyers depends on whether manufacturers can meet both cost and performance requirements. Now that many RF applications are leaving the talking stage and entering the building stage, perhaps the performance requirements will be clearer. With the economy slowly improving, perhaps more companies will have the ability to take the steps necessary to bring prices down. **RF**

RF design

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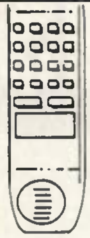

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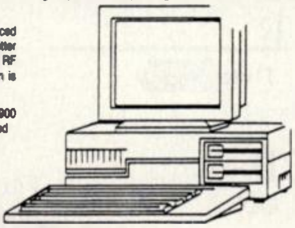

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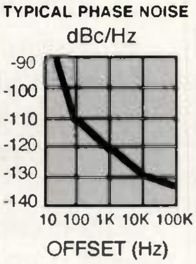
FREQUENCY SYNTHESIS

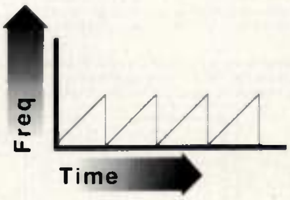
Bandwidth .... **1 to 400 MHz**
Step size .... <1 Hz
Frequency control .... 30-bits, parallel
Spurs .... <-45 dBc typ
Size .... 3.75" x 4.78" x 0.89"
Output .... **Quadrature**

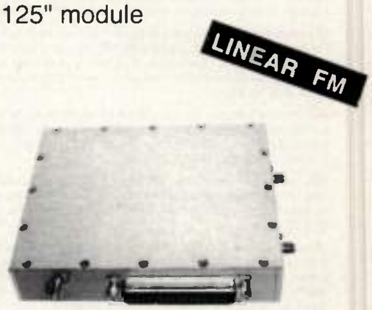


ADS-431
ADVANCED DDS

This **LOW COST** module is perfect for ATE, simulation, EW, and all other applications where speed and phase continuity are important.

Bandwidth .... **1 to 230 MHz**
Step size .... <30 Hz granularity
Phase control 12-bit
Switching .... 2 nanosecond update
Spurs .... -55 dBc typical (CW)
5.25" chassis or 5" x 7" x 1.125" module





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GaAs DDS

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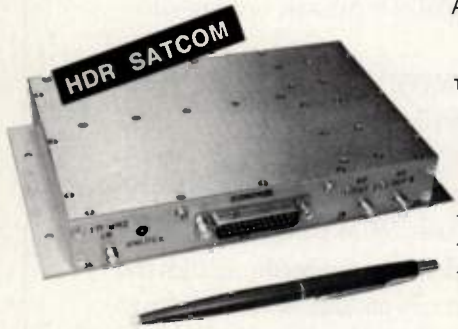
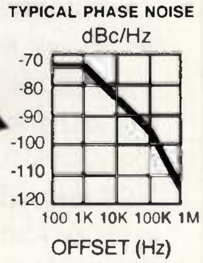
specializes in advanced technology **frequency synthesis**, using direct digital, phase-locked-loop, Arithmetically Locked Loop, and mix/filter designs, plus unique combination architectures that combine the advantages of multiple underlying technologies.

The VDS-6030 is a low cost C-band solution, with excellent phase noise and low power dissipation. It's also available in L-band.

VDS-6030
ARITHMETICALLY LOCKED LOOP

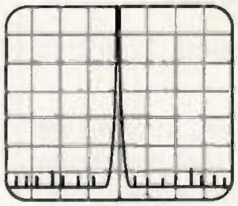




Bandwidth .... **4.65 to 5.25 GHz**
Step size .... 2.5 MHz
Outputs .... 2-channels @ +17 dBm
Total power .... <5 W
BIT .... TTL lock indicator
Spurs .... -60 dBc
Reference .... 10 MHz, ext (internal optional)

This one square inch waveform generator includes digital phase, frequency, and amplitude control, yet it only dissipates 1.5W at f_{ck} max.

DDS-1
SYNTHESIZER & MODULATOR


TYPICAL SPURS

Maximum clock .... **up to 25 MHz**
Frequency control 32-bit
Switching speed .... <1 µsec
Output .... SINE, 1 VP-P
Phase noise .... per the clock
Gating .... >>100 dB ON/OFF, 1-bit toggle
Modulation ...... **digital amplitude, ϕ, FSK**

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