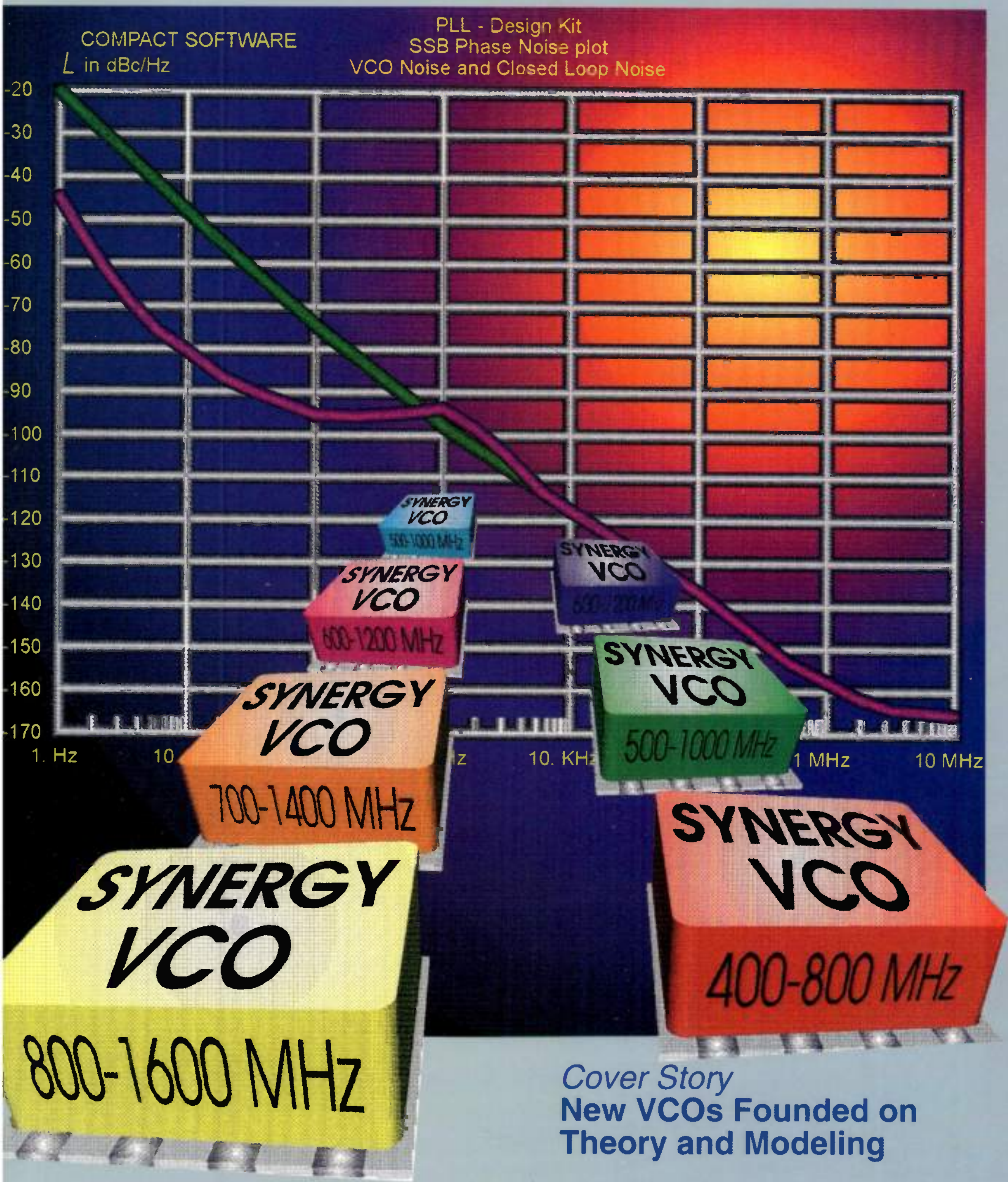
RF design

engineering principles and practices

June 1993



Cover Story

New VCOs Founded on Theory and Modeling

Featured Technology

Broadcasting Applications

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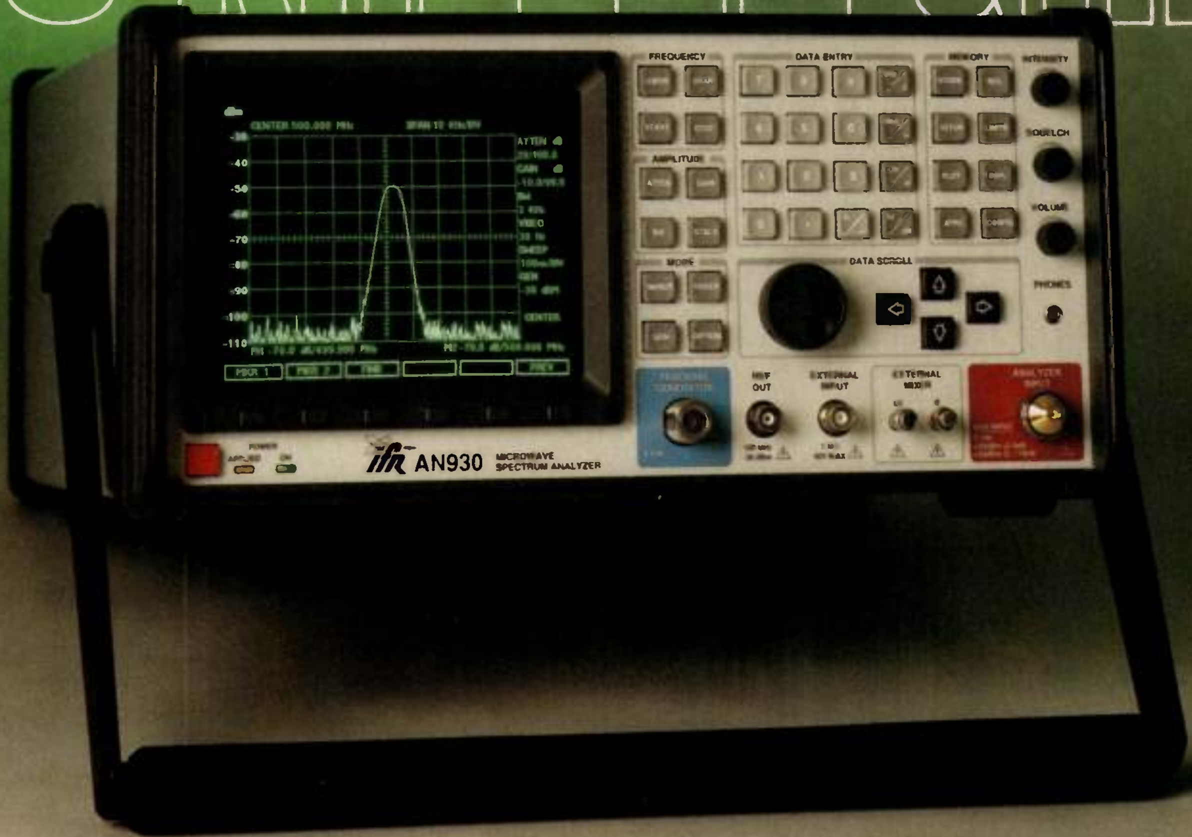

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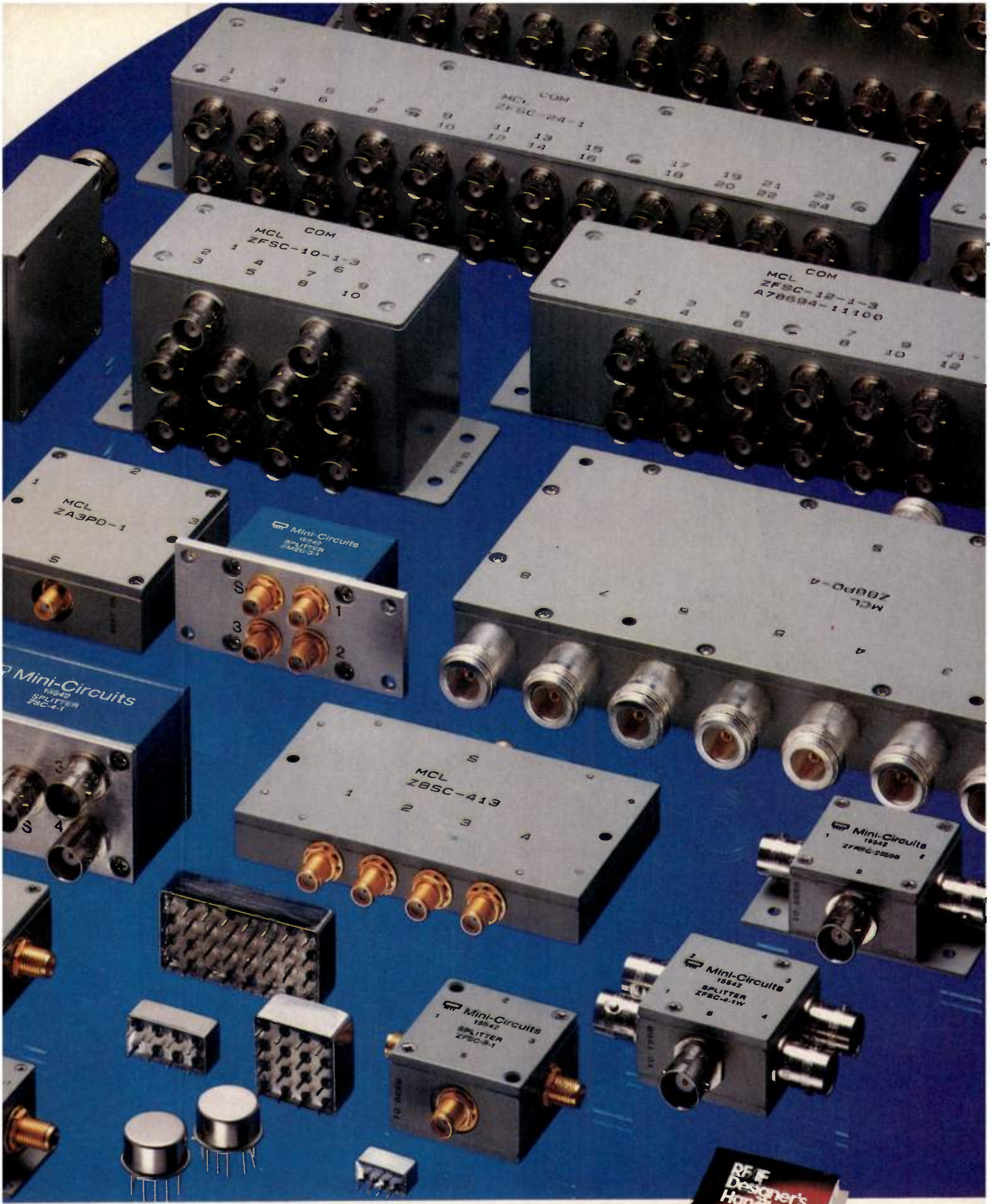
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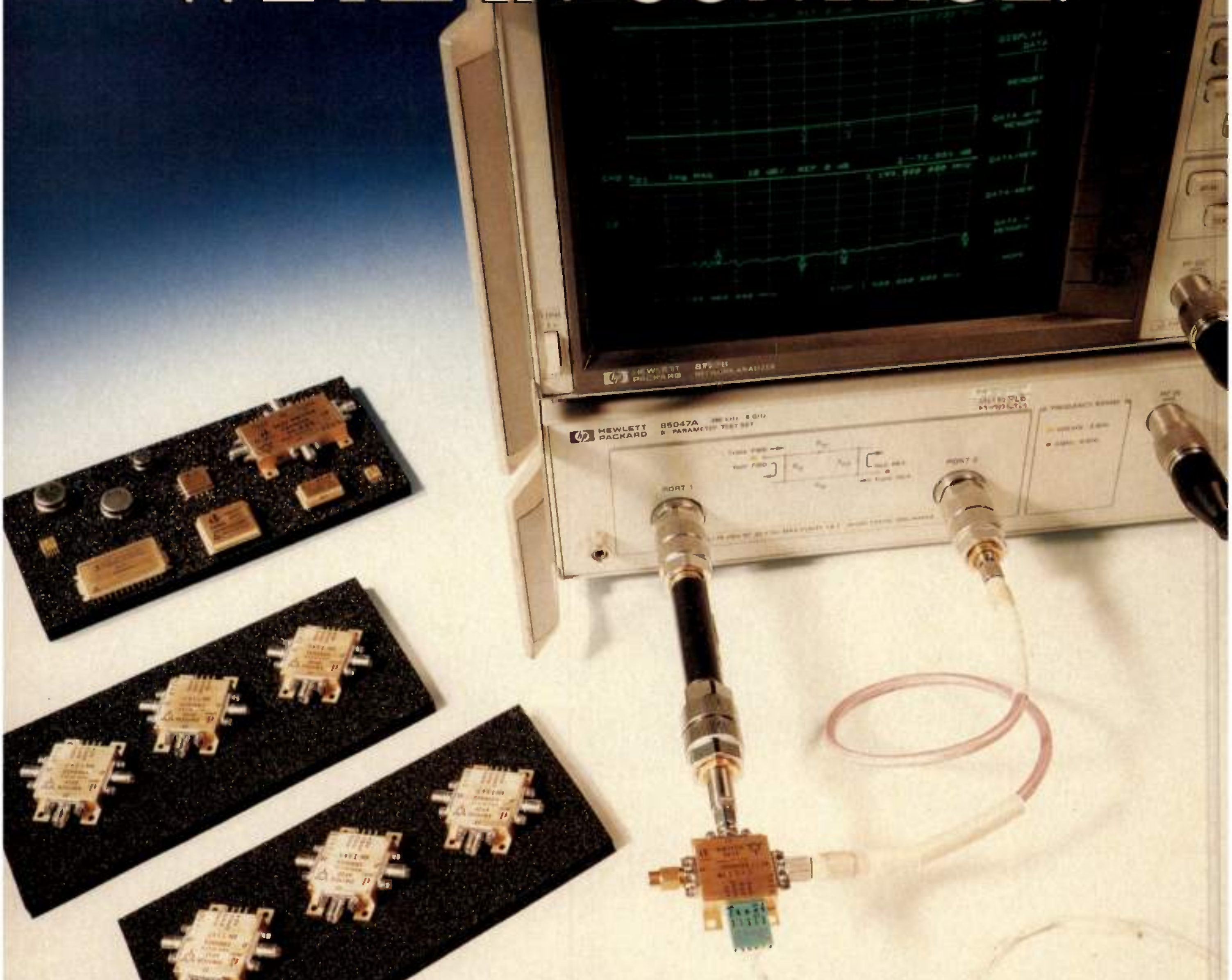

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SP2T	20-2000	0.7	85	3.000	TTL	SMA	CDSO622
SP2T	5-4000	1.0	79	0.035	TTL	SMA	CDSO882
SP4T	20-2000	0.9	72	3.000	TTL	SMA	CDSO624
SP2T	10-2000	0.60	82	1.000	TTL	14 Pin DIP	DSO052
SPST	10-2000	1.8	67	0.035	TTL	14 Pin SMP	DSO790
SP2T	DC-2000	0.7	50	0.200	TTL	TO-5	DSO813-T
SP4T	DC-2000	1.7	70	0.075	TTL	14 Pin DIP	DSO874

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# Of Sections	Freq MHz	LSB Range dB	IL dB	Switch Speed μ SEC/ Max	Control	Package	Part No.
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7	DC-1000	0.5-63.5	7.1	0.100	TTL	38 Pin DIP	DAO897
7	10-1000	0.5-63.5	6.4	0.035	TTL	SMA	CDAO867

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RF design

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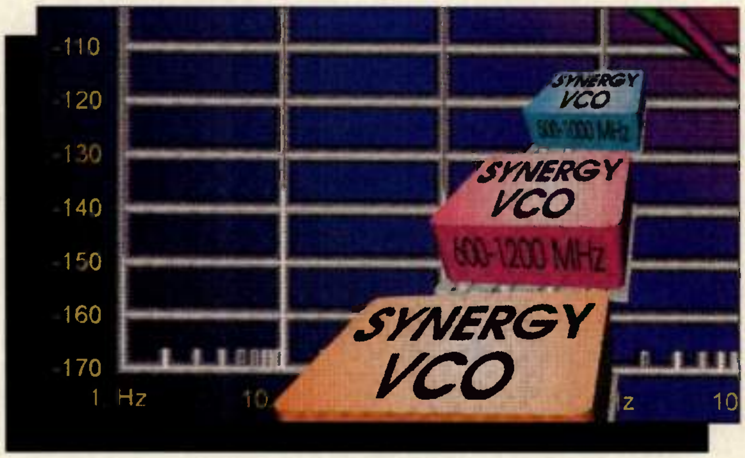
June 1993



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RF editorial

What is "Wireless?"

By Gary A. Breed
Editor

Everyone has a different definition of the term *wireless*! The mobile communications industry considers it to mean just cellular, mobile data and PCN/PCS. The computer industry only acknowledges wireless LANs, radio-linked peripherals, and perhaps cellular hookups used with "personal digital assistants." The automotive industry looks at wireless as vehicle locating systems, keyless entry and alarm systems. Other parts of the transportation industry see only vehicle tracking systems and automatic toll collection.

With so many opinions on the meaning of wireless, it's no wonder that the general public hasn't a clue about the size and scope of new RF product development. Yet, we expect these consumers to make a decision to buy new products using wireless technology. Of course, I have heard the argument that the radio portions of most new wireless products are supposed to be invisible to the consumer; and, therefore, the consumer does not need to know anything about the technology. This may be true when an application becomes a mass market item, but the *first* buyers of new technology are educated and aware. These trend-setters need to know what's inside, and they need some knowledge of capabilities and limitations.

I see the same lack of understanding among business and financial analysts. They form their opinions based on their industry contacts — and they have heard the same limited views of wireless that I pointed out at the start of this column. This is not good news! We want investors in new RF technology to understand that the wireless renaissance represents a huge market. We don't want them to think in terms of modest-sized single markets like wireless LANs or home security. We want the financial community to understand that RF is an enabling technology for many products — a total market that is on the same scale as the personal computer boom.

What can you do to raise awareness? First, you can make sure that your own companies understand the entire scope of new applications. If you are at the staff level, make sure your conversations with colleagues and supervisors include comments about the "big picture." Seemingly routine conversations with your banker, tax accountant or other friends and colleagues can plant seeds of information.

If you are a higher-ranking engineer or manager, you can really make an impact. Simple statements can present a subtle message, like, "RF is really exciting right now," or, "There are so many things happening in RF that I can't keep track of them all." Comments like this attract questions. Be ready to respond with explanations about RF technology and the many products being readied for the marketplace.

Each of us will benefit from growth in the RF marketplace. And if we all help promote the new ideas now taking shape, that growth will happen sooner.

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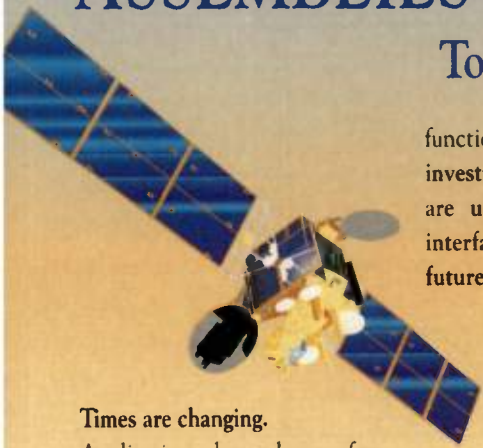

Times are changing.
Applications demand more from components, subassemblies and subsystems. The age of componets is evolving to an age of signal interfaces: Analog to Digital, RF to Microwave, Microwave to Microwave and Microwave to Optic. For thirty eight years, Sage has developed technology to meet the most difficult component performance requirements. Our technology spans electromechanical products, connectorized passive components, and patented caseless Wireline™ hybrid and coupler products. Having such diverse interface technology creates design flexibility and permits Sage to integrate component requirements into signal processing functions. **At Sage, major investments in new technology are ushering in products and interfaces for the future, a future era we call the Sage Age.**

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INTERFACE PRODUCT CAPABILITY	FREQUENCY RANGES COVERED	CATALOG	CUSTOM COMMERCIAL DEFENSE	SPACE QUALIFIED
Electromechanical				
Switches	DC-18 GHz	✓	✓	✓
Phase Shifters	DC-26 GHz	✓	✓	✓
Rotary Joints	DC-40 GHz	✓	✓	✓
Passive Connectorized				
Filters	100 MHz - 26 GHz	✓	✓	✓
Hybrids	60 MHz - 26 GHz	✓	✓	✓
Couplers	60 MHz - 18 GHz	✓	✓	✓
Power Dividers	100 MHz - 18 GHz	✓	✓	✓
Drop-In Caseless				
Wireline Hybrids	25 MHz - 18 GHz	✓	✓	✓
Wirepac Hybrids	25 MHz - 4 GHz	✓	✓	✓
New Technology Products				
Interface Assemblies	DC-26 GHz		✓	✓

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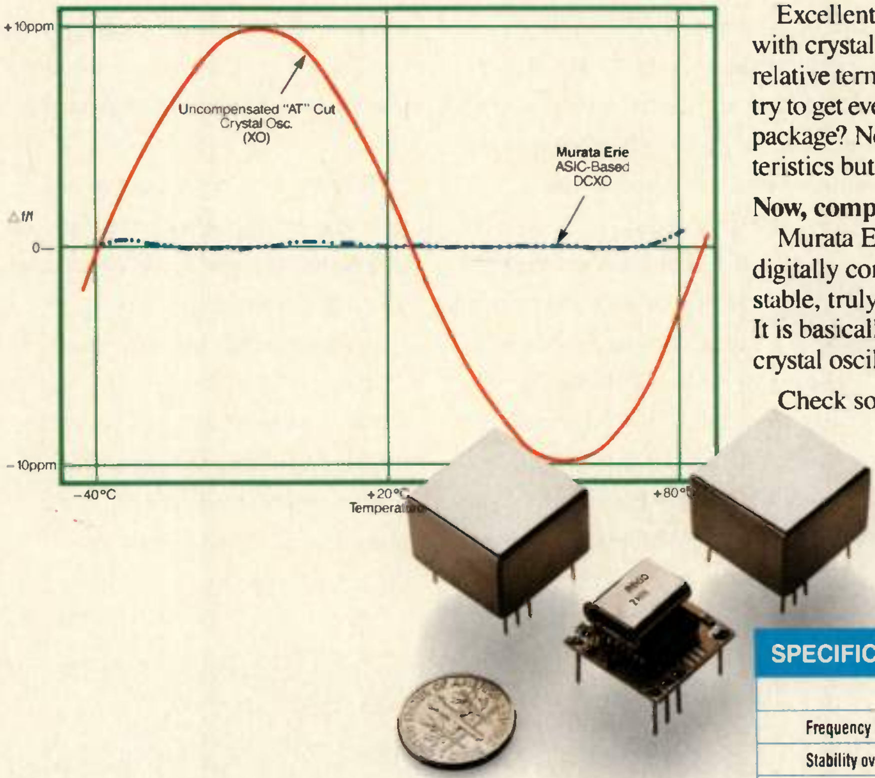
Murata Erie's new ASIC-based DCXO



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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SPECIFICATIONS

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Stability over −40°C to +85°C	±0.5ppm
Output	"HC" CMOS
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Short Term Stability	1×10E−9 @ T=1 sec.
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For the very first time, you can make CW *and* peak power measurements quickly and accurately with a single meter—a Universal Power Meter.

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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

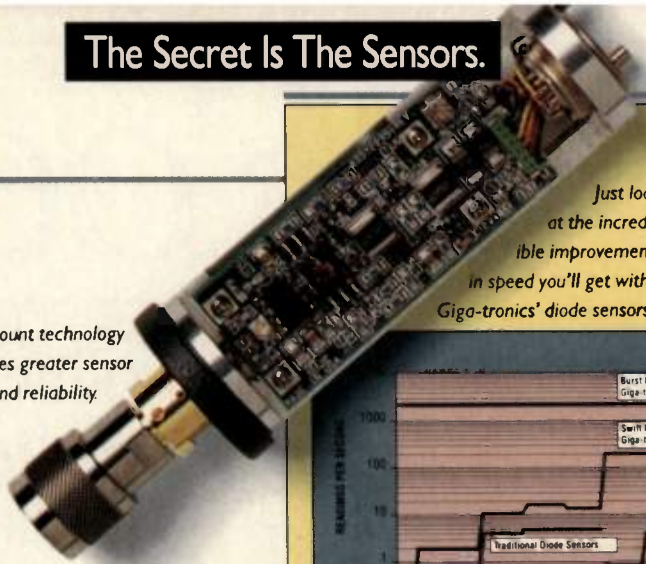
Now, an easy-to-use CW power meter can also measure pulsed RF signals with the simple addition of a peak power sensor.

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.

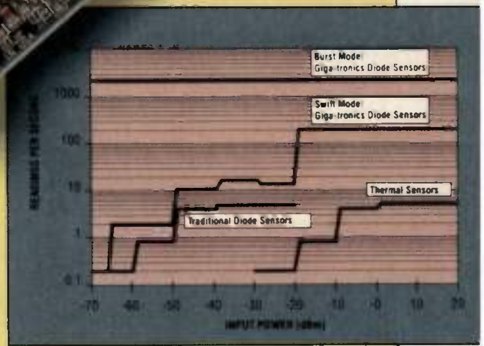
The 8540 Series features intelligent design and sophisticated software. The result is a simplified front panel and extensive built-in capabilities that prevent many common operational mistakes that typically lead to inaccurate measurements.

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 TRUE 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

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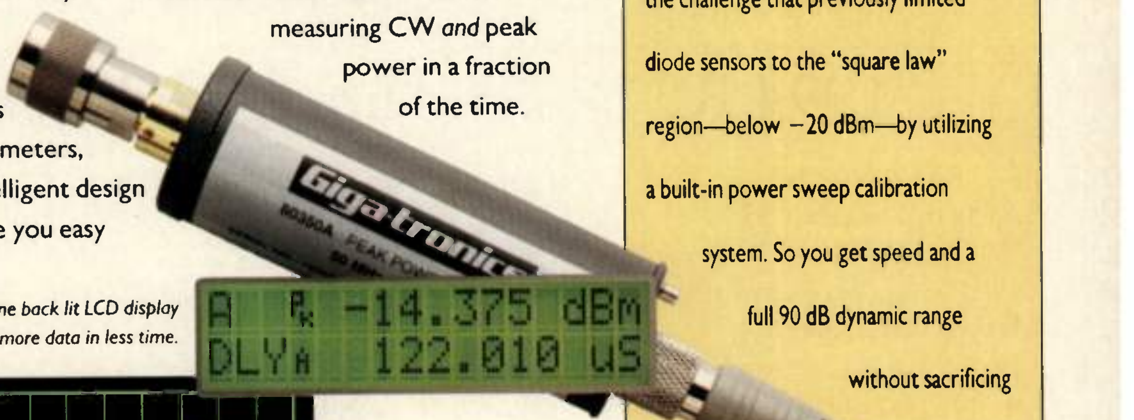
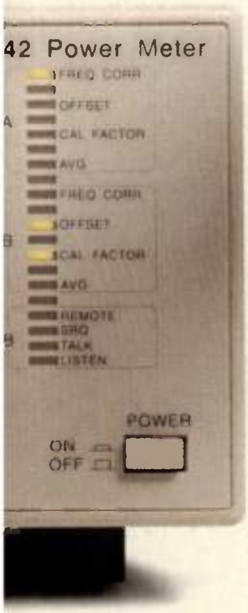
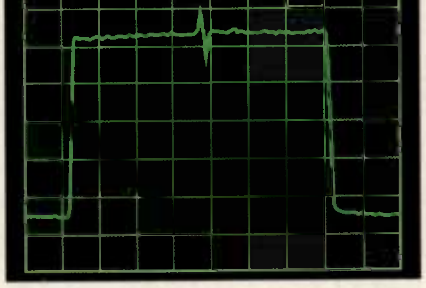


A two-line back lit LCD display provides you more data in less time.





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

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STEL-1173RH	40 MHz	Rad Hard with 1 MRad Tolerance
STEL-1174	50 MHz	16-bit NCO, Low Price
STEL-1175	80 MHz	32-bit NCO with Linear PM
STEL-1176	80 MHz	83/4 Decade Decimal NCO with BCD Control
STEL-1177	60 MHz	32-bit NCO with Linear PM and FM ports
STEL-1178A	80 MHz	32-bit NCO Dual with PSK
STEL-1179	25 MHz	24-bit NCO, PSK and Low Price
STEL-1180	60 MHz	32-bit Chirp Generating NCO
STEL-2172	300 MHz	28-bit ECL NCO
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BOARD LEVEL DDS PRODUCTS

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STEL-1273	0 to 22 MHz	DDS with Sub- MicroHz Resolution
STEL-1275	0 to 35 MHz	DDS with Linear PM
STEL-1276	0 to 35 MHz	DDS with 0.1 Hz Resolution and BCD Control
STEL-1277	0 to 35 MHz	DDS with Linear PM and FM
STEL-1375A	0 to 35 MHz	Miniature DDS Module with Linear PM
STEL-1376	0 to 35 MHz	Miniature DDS Module with BCD Control
STEL-1377	0 to 35 MHz	Miniature DDS Module with Linear PM and FM
STEL-1378A	Dual 0 to 35 MHz	Miniature DDS Module with PSK
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STEL-2272	0 to 130 MHz	DDS
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DENMARK: Vallentin Elctronik ApS, Mr. Lars Vallentin, Phone: +45 42 17 24 17, FAX: +45 42 17 15 18

JAPAN: Marubun Corporation, Mr. Toshi Ishizawa, Phone: +81 3 3639-9821, FAX: +81 3 3661-7433

INDIA: Accutrol Systems, Pvt., Ltd., Mr. Vivek Raghavan, Phone: +91 812 648079, FAX: +91 22 202-9403

ISREAL: Regev Aviation, Mr. Zeev Regev, Phone: +972 3 533-4359, FAX: +972 3 533-9302

SWEDEN: I. E. K., Mr. Bill Osterlund, Phone: +46 8 80 46 85, FAX: +46 8 26 22 86

AUSTRALIA: Sabtek Electronics, Mr. Grant Amor, Phone: +61 8 373-0233, FAX: +61 8 373-0206

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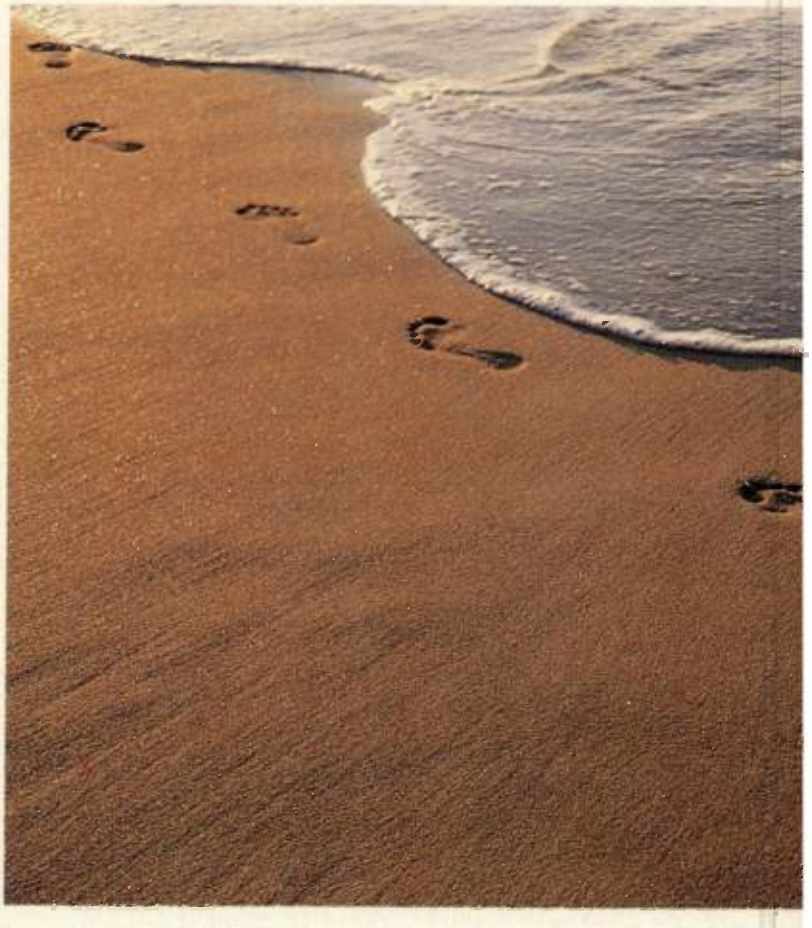

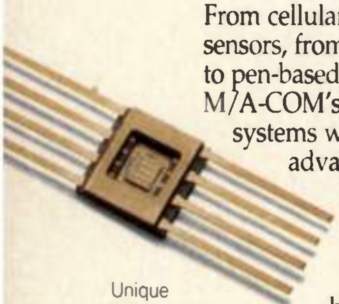

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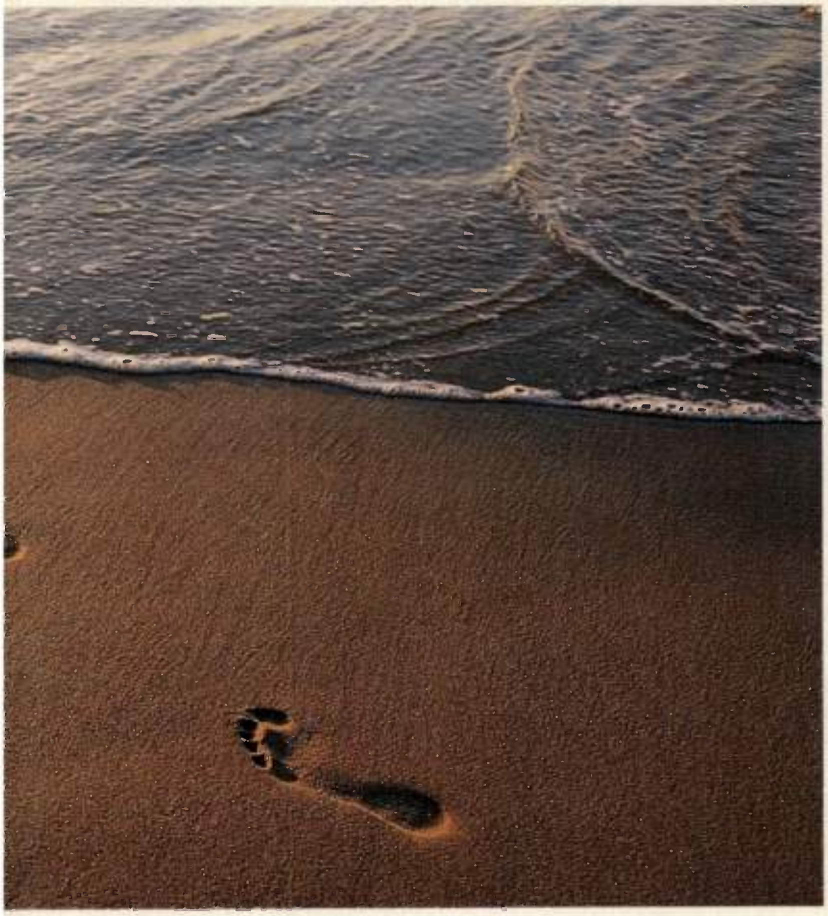

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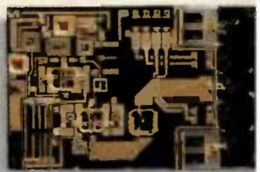

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RF letters

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.

A Second Look at Double Tuned Circuits

Editor:

Following the RF Tutorial, "Double Tuned Circuits", by Andrzej B. Przedpelski, (*RF Design*, Nov. 1992), I want to make several additional remarks.

Regardless of the circuit topology, the coupling coeffcient can be computed starting from the two possible coupling configurations presented in Figure 1a and 1b, where all the reactances are of the same type (capacitive or inductive).

For the circuit of Figure 1a, the coupling coefficient has the expression:

$$k_c = X_c / \sqrt{(X_{01} + X_c) \cdot (X_{02} + X_c)} \qquad (1)$$

while for the circuit of Figure 1b, the coupling coefficient is:

$$k_c = \sqrt{\frac{X_{01} \cdot X_{02}}{(X_{01} + X_c) \cdot (X_{02} + X_c)}} \qquad (2)$$

The employment of equations 1 and 2 for different double tuned circuit configurations leads to well known formulas. Even "strange" topologies, as that of Figure 1c, can be handled with equation 1 considering that L_{02} is in series with $C_{02} = \infty$ F. The coupling coefficient is

$$k_c = \sqrt{C_{01}/C_C}$$

I also consider it interesting to notice that if the tuned circuits are mutally coupled as in Figure 2b, then increasing or decreasing k_C has no effect on the resonant frequency of the two separate circuits, (and hence on the central frequency of the double-tuned circuit characteristic), and the frequency characteristic for different coupling coefficients has the same shape as in Figure 3.

If capacitive or inductive (not mutual!) coupling is employed, then the situation changes. For example, in the double-tuned circuit from Figure 2a, as C_C is increased (and consequently the coupling is increased), the resonant frequency of both circuits decreases (which is absolutely normal, because C_{01} and C_C are somehow in parallel). The global effect is that the selective characteristic tends to leave almost unchanged its right peak position and shifts to the left its left peak, as presented in Figure 4.

Ion-Constantin Tesu
Iasi Poltechnic Institute, Romania

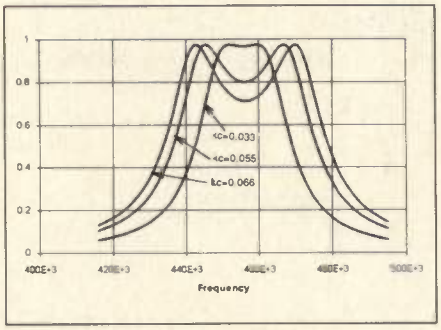

Figure 3.

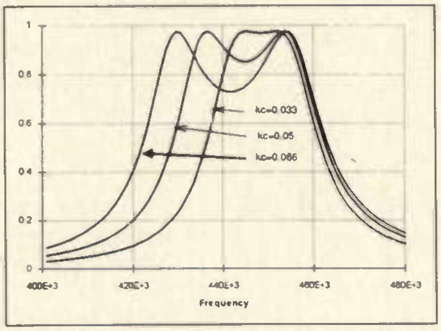

Figure 4.

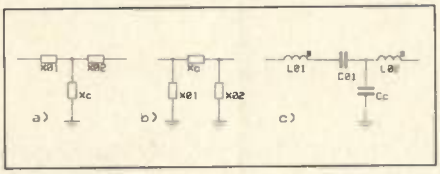

Figure 1.

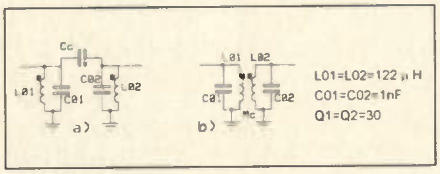

Figure 2.

DECT IC Info

In the April Cover Story, "Radios for the Future: Designing for DECT," we somehow included an INFO/CARD number that was not on our reply card. We apologize for the mistake. Readers interested in the National Semiconductor DECT chipset described in that article can get more information by calling National's Customer Service Center at (800) 272-9959.

RF calendar

June

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.

15-17 **Electric 93**
Atlanta, GA
Information: Continental Exhibitions, Inc., 370 Lexington Avenue, Suite 902, New York, NY 10017-6578. Tel: (800) 222-2596. Fax: (212) 370-5699.

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.

27-2 **1993 IEEE AP-S International Symposium and URSI Radio Science Meeting**
Ann Arbor, MI
Information: University of Michigan Conferences and Seminars, Ms. Ann Pendleton. Tel: (313) 936-0379. Fax: (313) 764-2990.

28-30 **Spread-Spectrum Communication Systems & Applications**
Ann Arbor, MI
Information: Engineering Conferences, 400 Chrysler Center, North Campus, The University of Michigan, Ann Arbor, MI 48109-2092. Tel: (313) 764-8490. Fax: (313) 936-0253.

August

2-5 **International Microwave Conference**
Sao Paulo, Brazil
Information: Paulina Cardoso, IMT-Escola de Engenharia Maus, Estrada das Lagrimas, 2035, 09580 S. Caetano do Sul - SP, Brazil.

9-13 **IEEE International Symposium on Electromagnetic Compatibility**
Dallas, TX
Information: Dr. Frederich M. Tesche, c/o International Compliance Corporation, 1911 E. Jeter Rd., Argyle, TX 76226. Tel: (817) 491- 3696. Fax: (817) 491-3699.

18-21 **IEEE 1993 International Geoscience and Remote Sensing Symposium (IGARSS '93)**
Tokyo, Japan
Information: Mr. Natsuhiko Motomura, Remote Sensing Technology Center of Japan, 7-15-17 Roppongi, Minato-ku, Tokyo 106, Japan. Tel: (81) 3-3403-1761. Fax: (81) 3-3403-1766.

RF courses

Basic Radar
July 13-15, 1993, Atlanta, GA
Radar Design Workshop
August 17-19, 1993, Atlanta, GA
Information: Georgia Institute of Technology, Continuing Education. Tel: (404) 894-2547.

Wavelet Transform: Techniques and Applications
August 9-11, 1993, Los Angeles, CA
Active and Passive RF Components: Measurements, Models, and Data Extraction
August 16-20, 1993, Los Angeles, CA
Information: UCLA Short Course Program Office. Tel: (310) 825-1047. Fax: (310) 206-2815.

High-Speed Communication Networks
August 16-18, 1993, Santa Cruz, CA
Compression Technologies: Image, Video & Associated Standards for Computers, Communications & Consumers
August 19-20, 1993, Santa Cruz, CA
Information: University of California Extension. Tel: (408) 427-6600. Fax: (408) 427-6608.

Analyzing Communication System Performance
July 14-16, 1993, Washington, DC
September 13-15, 1993, San Diego, CA
Digital Cellular Radio
July 27-30, 1993, Washington, DC
Nonlinear Microwave Circuits
August 9-11, 1993, Washington, DC
Telecommunication Traffic Engineering
August 16-18, 1993, Washington, DC
Modern Receiver Design
August 23-27, 1993, Washington, DC
Microwave Radio Systems
August 25-27, 1993, Washington, DC
Introduction to Wireless Telecommunications: Technologies, Applications, Regulatory Issues, and Market Dynamics
August 30-September 1, 1993, Washington, DC
Mobile Satellite Communication Systems
August 30-September 1, 1993, Washington, DC
Principles and Applications of High-Frequency Radio Communications
August 30-September 2, 1993, Washington, DC
Lightning Protection
September 9-10, 1993, Washington, DC
Global Positioning System: Principles and Practice
September 15-17, 1993, San Diego, CA
Future Telecommunications for Providers, Suppliers, Users, and Regulators
September 20-22, 1993, Washington, DC
Information: The George Washington University, Continuing Engineering Education, Merril A. Ferber. Tel: (202) 994-8522 or (800) 424-9773.

Workshop in Finite Elements in Electromagnetics
June 21-25, 1993, Troy, NY
Information: Rensselaer Polytechnic Institute, Office of Continuing Education. Tel: (518) 276-8351.

Navstar/GPS
September 22-24, 1993, Boston, MA
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 Systems
July 18-23, 1993, Guildford, UK
Information: The Institution of Electrical Engineers, Savoy Place, London WC2R 0BL, United Kingdom. Tel: (44) 071-240 1871. Fax: (44) 071-497 3633.

Electromagnetic Compatibility and Interference
September 14-17. 1993, San Diego, 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.

Linear Design Seminar
July 14, 1993, Raleigh, NC
July 15, 1993, Atlanta, GA
July 27, 1993, Dayton, OH
July 28, 1993, Cleveland, OH
July 29, 1993, Philadelphia, PA
August 11, 1993, Detroit, MI
August 12, 1993, Chicago, IL
August 17, 1993, Iselin, NJ
August 19, 1993, Orlando, FL
Information: Texas Instruments. Tel: (800) 477-8924 x3443.

RF/MW Small Signal/Low Noise Amplifier Design
June 13-14, 1993, Atlanta, GA
RF/MW Large Signal Amplifier Design
June 13-14, 1993, Atlanta, GA
Applied RF Design Techniques I
June 21-25, 1993, Burlington, MA
RF/MW Circuit Design
July 5-9, 1993, Oxford, UK
RF Circuit Components: Measurements, Models and Data Extraction
July 12-16, 1993, Okford, UK
August 16-20, 1993, Los Angeles, CA
RF Circuit Design: Passive and Active Linear Networks
August 24-27, 1993, Los Altos, CA
RF Design: Nonlinear Circuits and Devices
August 30-Sept 2, 1993, Los Altos, CA
Information: Besser Associates. Tel: (415) 949-3300. Fax: (415) 949-4400.

Inherently Conductive Polymers: An Emerging Technology
September 8-10, 1993, Boston, MA
Information: Advanced Polymer Courses, Dr. M. Aldissi. Tel: (802) 655-2121. Fax: (802) 655-2025.

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	210LC	10W CW	.008-225 MHz	40dB	$ 2,495
	710FC	10W CW	1-1000 MHz	40dB	$ 6,695
	*727LC	10W CW	.006-1000 MHz	44dB	$ 7,750
NEW	713FC	15W CW	10-1000 MHz	42dB	$ 4,250
	225LC	25W CW	.01-225 MHz	40dB	$ 3,295
	*737LC	25W CW	.01-1000 MHz	45dB	$ 9,995
	712FC	25W CW	200-1000 MHz	45dB	$ 6,950
NEW	714FC	30W CW	10-1000 MHz	45dB	$ 9,950
	250LC	50W CW	.01-225 MHz	47dB	$ 5,250
	715FC	50W CW	200-1000 MHz	47dB	$ 16,990
	707FC	50W CW	400-1000 MHz	50dB	$ 9,990
NEW	716FC	50W CW	10-1000 MHz	47dB	$ 17,950
	*747LC	50W CW	.01-1000 MHz	47dB	$ 19,500
	116FC	100W CW	.01-225 MHz	50dB	$ 8,800
	709FC	100W CW	500-1000 MHz	50dB	$ 16,990
	717FC	100W CW	200-1000 MHz	50dB	$ 19,500
NEW	718FC	100W CW	10-1000 MHz	50dB	$ 26,950
	*757LC	100W CW	.01-1000 MHz	50dB	$ 29,950
	122FC	250W CW	.01-225 MHz	55dB	$ 14,650
	723FC	300W CW	500-1000 MHz	55dB	$ 29,995
	LA500G	500W CW	500-1000 MHz	57dB	$ 53,500

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122C	200W CW	.01-220 MHz	53dB	$ 11,750
134C	500W CW	.01-220 MHz	57dB	$ 19,800
137C	1000W CW	.01-220 MHz	60dB	$ 27,550
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ITU Streamlines Telecommunications Sector

The World Telecommunication Standardization Conference recently completed a meeting to streamline the Telecommunication Standardization Sector to increase its competitiveness in the global standardization arena. An advisory group was set up to review priorities and strategies for activities of the Telecommunication Standardization Sector, to review progress in the implementation of its work program and to recommend measures to foster cooperation and coordination with other standards bodies and with other organizations having interests in telecommunication standardization. In addition, the Conference officially launched the ITUDOC — the ITU's electronic document exchange service. The service includes administrative, general information documents, and ITU standards. The Conference also announced the availability of all ITU telecommunication standards on CD-ROM.

NIST Budget Increased — President Clinton's fiscal year 1994 budget request of $535.2 million for the Commerce Department's National Institute of Standards and Technology is a $151.2 million, or 39 percent, increase over the current appropriation of $384 million. $241 million will be spent to fund research and supporting services including initiatives in high performance computing, advanced manufacturing, advanced materials, electronics, biotechnology and chemical processing and international trade and standards. $232.5 million has been appropriated for technology development and technology transfer and $61.7 million will fund design work for a 10-year project to replace scientifically obsolete laboratory space at NIST's 25- to 35-year-old facilities in both Maryland and Colorado.

U.S. and Russia to Harmonize Standards — A memorandum of understanding on scientific and technical cooperation was recently signed by the Department of Commerce's National Institute of Standards and Technology and the State Committee of the Russian Federation for Standardization, Metrology and Certification. Both parties will work together to promote international standards and product acceptance criteria; develop new methods and reference standards/materials for different types of measurements; harmonize standards for legal metrology; research precise measurements of physical quantities and comparisons of standards of basic physical units; perform fundamental research in chemical, physical and engineering metrology; and other activities related to standards and/or metrology.

New Passivating Process for Gallium Arsenide — Researchers at the Georgia Institute of Technology have developed a new process for stabilizing gallium arsenide semiconductors. The process protects the gallium arsenide device from corrosion, mechanical damage and electrical deterioration. The new technique produces a stable nitride film at low temperatures and does not damage the fragile crystalline structure of the material. The technique can be

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used with other III-V compound semiconductors as well. Georgia Tech scientists believe the nitriding process could be used to insulate and seal the junctions of capacitors, to insulate areas between transistors, and to provide the dielectric layer for making capacitors. Because it uses low temperatures, the process could be used to passivate a fully-fabricated electronic device.

Multi-GHz Bandwidth Superconducting ADC — Hypres, Inc. recently developed and tested a 10-GHz bandwidth superconducting flash analog-to-digital converter (ADC) that achieves 4.4 effective number of bits (ENOB) at a 4-GHz signal bandwidth. An 8-bit device was tested in liquid helium at a temperature of 4.2K. Tests includec sampling low-bandwidth (100 kHz) and high-bandwidth (2 to 12 GHz) signals at a 1-Gsample/s rate. At 100 kHz, the ENOB was measured to be 7.5; at the higher signal range it was 7.25. The ADC was fabricated with 2-um junctions at 5000A/cm^2 using Hypre's standard eight-layer niobium process. Characteristics include low circuit complexity, less than 1 mW power consumption, and a full scale sensitivity of 10 mA into 25 ohms.

Commercial Satellite Equipment Market to Grow — According to a recent report from Market Intelligence entitled, "Satellite Communications Equipment Markets: VSAT and Mobile Applications Unleash Satcom," the commercial satellite equipment market is expected to expand from $1.3 billion worldwide in 1992 to $2 billion in 1998 at a 7 percent compound annual rate. Major end-user growth in demand will come from retail, automotive and media businesses. Along with private business networks, demand for distance learning and public sector applications — especially in less developed parts of the world — will fuel earth station growth. According to the report, most growth will take place outside North America.

Siemens Division Relocates — The Siemens Integrated Circuit Division recently relocated. Their new address is: 01950 North Tantau Avenue, Cupertino, CA 95014. Tel: (408) 777-4500, Fax: (408) 777-4957.

Samsung Acquires Interest in array Microsystems — array Microsystems has announced that Samsung Electronics Company has acquired a 20 percent interest in array Microsystems through an equity investment. Samsung and array have been working together since January 1991 on a family of video compression ICs targeted at the multimedia market.

Oscillator Markets, Applications and Competitors — A new market survey is available from Allied Business Intelligence entitled, "Oscillators: North American Markets, Applications & Competitors: 1993 to 1996 Analysis." The report looks at production, market shares, trends, competition, technology, forecasts and applications. For example, according to the report 55 percent of oscillators sold to the merchant market in North America in 1992 were for narrowband applications. Growth is expected to be around 7.25 percent annually through 1996 for these products.

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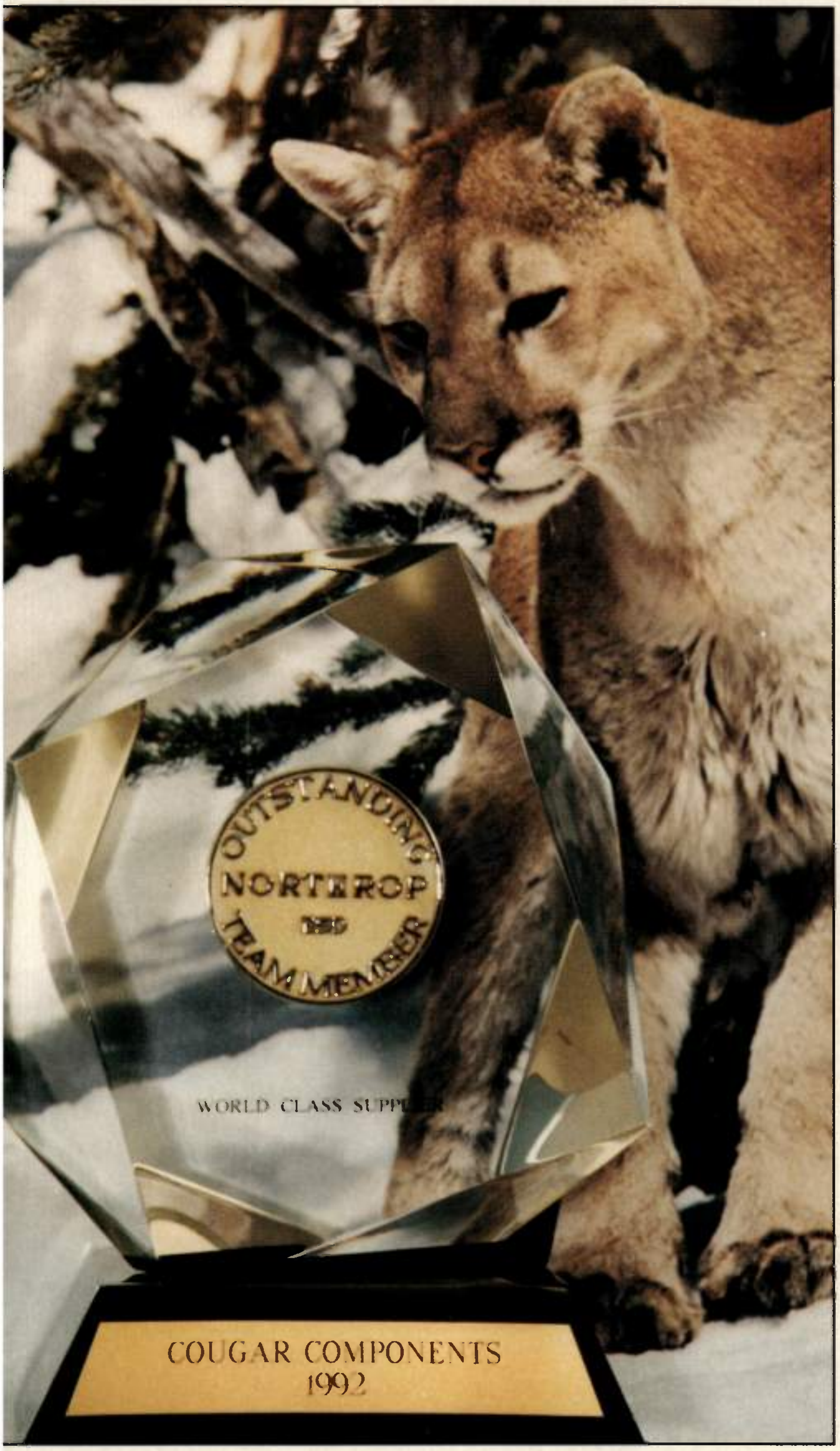

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Broadcast Symposium Call for Papers — The 43rd Annual Broadcast Engineering Symposium sponsored by the Broadcast Technology Society of the IEEE has issued a call for papers. The symposium will be held September 22-23, 1993 in Washington, DC. Technical papers should focus on AM, FM and Television digital transmission techniques; antenna design and testing; transmitter development and other related topics. Send abstracts by June 15 to Philip A. Rubin, Rubin Bednarek & Associates, 1350 Connecticut Avenue, NW #610, Washington, DC 20036. Fax: (202) 296-9383.

Disaster Relief Aided by New Cellular Technology — In December, 1992, the Indonesian Island of Flores was devastated by an earthquake and tidal wave that caused the deaths of more than 2000 people and destroyed much of the island's infrastructure. To aid relief efforts, Motorola implemented their new Wireless Local Loop product, WILL®. The system was developed to serve the basic telephony needs of people in urban and difficult to reach rural areas and is intended to provide fixed telephony services in areas with little or no existing wireline telephone service or as a supplement to the existing wireline service. From order placement to system optimization, the WILL® project took three weeks; installation of the operational system was completed in five days.

Microwave Hybrid Circuits Call for Papers — Rogers Corporation has issued a call for papers for the 1993 Microwave Hybrid Circuits Conference. The conference, to be held October 17-20 in Apache Junction, AZ, will address the theme: "Microwave Packaging for the 90's." New MIC/MMIC papers on microwave or millimeter work will be considered. Suggested subjects include: new applications for microwave; innovative design, processing and quality issues; advancement in materials; materials testing; packaging techniques; unsolved problems in design or implementation; software tools; and high volume manufacturing methods. Papers which address the theme of this year's conference will be given special consideration. Send abstracts as soon as possible to: General Chairman, Rogers Corp., 100 S. Roosevelt Ave., Chandler, AZ 85226. Tel: (602) 961-1382. Fax: (602) 961-4533.

Compression Labs to Help Develop Digital Simulcast HDTV System — Compression Labs recently announced that they will participate in the development of the Advanced Television Research Consortium's (ATRC) digital simulcast HDTV system. Their partners will include the David Sarnoff Research Center, Philips Consumer Electronics, Thomson Consumer Electronics and NBC. The system will undergo testing from the Federal Communications Commission's Advisory Committee System Test Program at the Advanced Television Test Center early next year.

RFID Open Protocol Released — Micron Communications, Inc. recently announced the publication of the "Micron RFID Protocol." The radio frequency identification protocol is open so that other companies can license it to make compatible RFI components and systems. Persons interested in receiving a copy of the protocol may contact Kae Wiemer at (208) 368-4000.

Dynascan Changes Name — Dynascan Corporation has announced a name change to Cobra Electronics Corporation. The change comes in response to the recognition the company has received for their Cobra brands which include cordless telephones, answering systems, CB radios, radar and laser detection systems and scanners.

Rapid Systems Moves to New Location — Rapid Systems, Inc. has announced the relocation of their offices to: 4307 Leary Way NW, Seattle, WA 98107. Tel: (206) 784-4311. Fax: (206) 784-0333.

Scientific-Atlanta Ships Inmarsat-M Terminals — Scientific-Atlanta has announced that its Model 9821 MariStar-M® is the first terminal to be approved by Inmarsat for connection to its new Inmarsat-M digital service network. Two yachts, the 92-foot *Ampro Viking* and the 46-foot *Jackerel* are now equipped with the capability to transmit and receive voice transmissions anywhere in the world, including the middle of the ocean. The unit consists of a lightweight above-deck antenna and a small below-deck electronics unit which is about the size of a VCR. Digital voice telephone service is provided at a compressed rate of 4.8 kilobits per second.

New EMC Company — Euro EMC Service recently announced their startup. The new firm offers EMC testhouse capabilities, consulting, training, and R&D. Their address is: Euro EMC Service (EES), Dr. Hansen GmbH, Potsdamer Str. 10 (TZT), D-O-1530 (14513) Teltow, Germany. Tel: (49) 3328 477141. Fax: (49) 3328 477142.

Consortium Formed to Advance Synthesizing, Processing of Electronic Materials — A team comprising five high-technology industrial firms, four universities and a national laboratory has undertaken a two-year program that promises to significantly advance the nation's capabilities in high-performance electronic and optical devices. The alliance, headed by Hughes Research Laboratories and Texas Instruments' Central Research Laboratory as principal partners, has been funded by the Advanced Research Projects Agency to study ways to advance the synthesis and processing of new electronic materials. The major goal of the program is to enhance molecular beam epitaxy technology, thus improving the U.S. competitive position in semiconductor structures. The program is valued at approximately $10 million with more than half the funding provided by the industrial partners.

Pineapple Technology Acquires Amplifier Product Line — Pineapple Technology Inc. has announced the acquisition of the amplifier product line developed for the TV broadcast industry from Spectrian (formerly MMD). The product line consists of solid-state amplifiers and subassemblies designed and manufactured for OEM's in the domestic and international TV broadcast markets.

Military RF Systems — Selecting the Right Targets

By Gary A. Breed
Editor

There is no question that the continuing reduction in military spending is wreaking havoc in many parts of the RF community. The threat of the Soviet superpower is probably gone for good, and although there remain other threats to U.S. interests worldwide, there is no more arms race. The cuts are deep enough to keep the most prominent Senators and Representatives from bringing lucrative military contracts to companies in their districts, and many bases are slated for closure.

Even with a smaller budget, however, the U.S. military is still a very large market for RF products. This report will highlight a few areas where significant military development and manufacturing is being funded.

Superconductivity

The increased performance and smaller size afforded by superconducting circuits has the attention of personnel directing military development funding. James Long, Vice President of Superconductor Technologies notes, "We should thank DARPA, which is now ARPA, for maintaining support for several key technologies." Recent projects at Superconductor Technologies include a notch filter bank for a program administered by Wright-Patterson AFB, a 35 GHz to 1 GHz downconverter, a 2 GHz miniature lumped-element filter, and several delay lines.

The military sees a benefit from superconductors in high performance microwave and high frequency circuitry. Low-loss printed delay lines can replace large, lossy coaxial cable assemblies without the high dispersion and radiation of ordinary micro-stripline. High performance filters can be constructed as loaded cavities or stripline instead of full-size cavities. The size and weight savings can be critical in airborne surveillance or EW systems.

The major problem area is in cooling the superconductors to liquid nitrogen temperatures. Fortunately, some high performance systems are already using cooled low-noise amplifiers, and can support additional system components. Superconductor Technologies is working closely with cryogenic cooler manufacturers to develop small, inexpensive and reliable coolers.

An additional benefit in the eyes of the funding authorities is transfer of technology to commercial markets. Superconductors are already in use as interference filters for radio astronomy, and in spacecraft communications and sensing equipment. Research is underway into low-loss MRI (magnetic resonance imaging) coils and cellular base station channel filters.

A soldier uses the AN/PRC-119 SINCGARS manpack radio.

Military Programs

"Communications systems are among the most active military programs right now," notes Roger Lesser, Editor of *Defense Electronics*. The SINCGARS manpack radio program (ITT Aerospace, prime contractor) and the MILSTAR satellite communications program (Lockheed, prime contractor) are both in full production. Although they have seen a slight slowdown in the delivery schedule, these programs have seen no cuts.

A more recent program noted by Lesser that offers a new opportunity is Combat Talon, a Special Operations support aircraft carrying communications, surveillance and countermeasures equipment. Loral, Hughes and Texas Instruments are mentioned as military contractors involved in this program. This type of program is an example of the military's interest in C^4I systems, which provide information that can make more efficient use of manpower and hardware. Also in this category are threat simulators, which allow training in a realistic environment, and which are also used for EW system evaluation. There is also continued work on smart munitions, which minimize pilot risk and improve combat effectiveness.

Even program cancellations can have a bright side. Recently, the Airborne Self-Protection Jammer (ASPJ) program was canceled, leaving the Navy without a replacement for older countermeasures systems. Numerous defense contractors (Loral, Lockheed Sanders, Litton, Raytheon, Tracor and Martin Marietta) are offering versions of existing systems that can update the Navy's F-18 and F-14 fighters at a lower cost than ASPJ, while keeping production and support personnel busy.

Of course, the NAVSTAR/GPS program must be noted. This military navigation system is rapidly becoming a significant civilian market, with extensive use in land surveying and commercial transportation. Hand-held GPS receivers are approaching the $500 price range. NAVSTAR/GPS is an outstanding example of a military program with a major spinoff into commercial and consumer markets. With budget cuts and careful spending, such multiple-use systems may have the edge in receiving funding for development. **RF**

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Implementation of a C-QUAM AM-Stereo Receiver Using a General Purpose DSP Device

By Dion Messer Funderburk and Sangil Park
Motorola Inc., Digital Signal Processing Operations

The audio consumer market has embraced digital audio systems with DSP (digital signal processing) as the latest and best technology for audio electronics. DSP is used in these systems to create concert hall effects, reverberation and noise cancellation as well as CD quality sound. Since these are all firmware solutions, the DSP's use can be extended to that of digitally receiving radio signals as well, which reduces analog circuitry in audio systems while providing superior quality sound.

The authors derived and implemented an algorithm to perform AM radio reception with and without the C-QUAM (C-QUAM is a registered trade mark of Motorola, Inc.) (Compatible Quadrature Amplitude Modulation) stereo signal. This paper outlines this derivation, analysis and implementation of the digital radio using digital signal processing techniques.

Background

The FCC approved a stereo system for FM broadcast at the peak of AM radio popularity. When this occurred in 1961, AM radio had the biggest share of the radio market (1). The stereo sound that FM provided quickly eroded the AM radio market and efforts were made to create a stereo broadcast system for AM. Many techniques were developed to perform AM stereo transmission to recover a larger share of the radio audience lost to FM.

To this day the FCC has not specified one particular AM stereo technique as a standard as they did with FM, but did complicate the design by specifying that every AM stereo broadcast technique must allow installed monaural AM receivers to receive an undistorted signal when tuned to a stereo channel. As much as this complicated the design of AM stereo techniques, the FCC tested and approved (approved, not specified) several standards including the C-QUAM technique in 1981 (2). There are more C-QUAM stations than any other AM stereo technique with approximately 574 stations throughout the United States and 206 stations internationally for a total of 780 C-QUAM stations world wide using the C-QUAM stereo transmission technique (3).

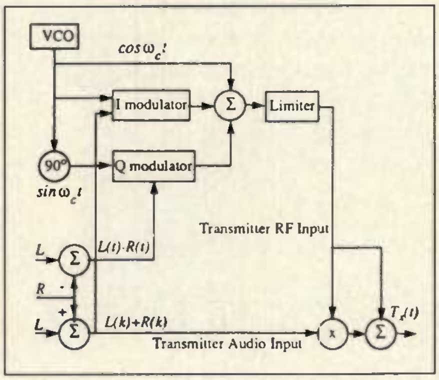

Figure 1. C-QUAM AM stereo transmitter.

The compatibility issue of receiving a stereo signal undistorted on a non-stereo receiver has made the most obvious method of providing an AM stereo signal impossible to use. The most obvious technique is that of modulating a carrier with the L + R signal and a 90 degrees shifted carrier with the L – R signal, generally referred to as Quadrature Amplitude Modulation (QAM). Envelope detectors are the most popular method of detecting the AM radio signals and they would not be able to detect this signal if a large modulation index existed in a stereo signal in which L only or R only was transmitted. Thus, it would not be compatible with existing receivers which use envelope detectors.

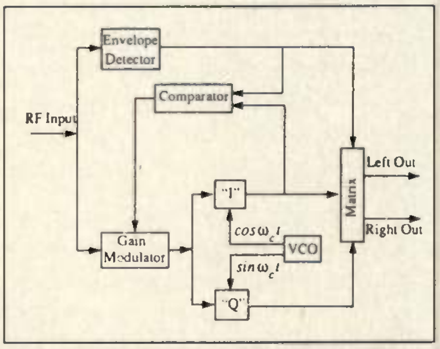

Figure 2. Analog C-QUAM AM stereo receiver.

A unique and patented method is used to create the C-QUAM stereo signal. This technique transmits L (left channel) and R (right channel) information by encoding L + R and L – R channel information into the phase of the transmitter carrier. Finally, the "modified" carrier is then amplitude modulated by

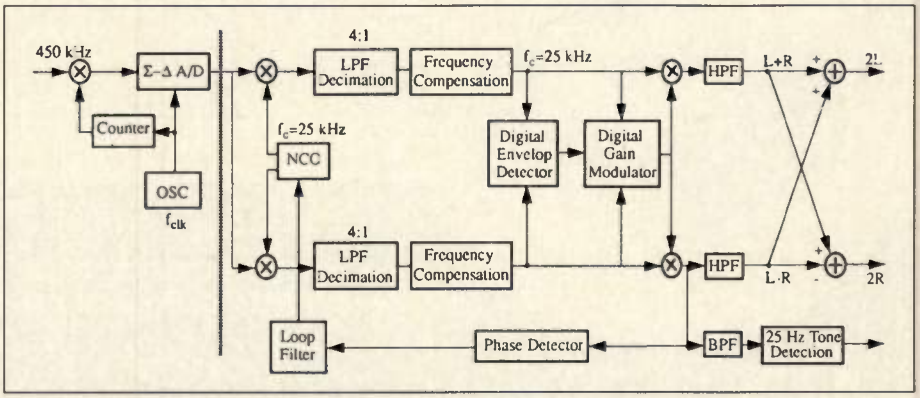

Figure 3. Detailed system diagram for the digital AM stereo receiver.

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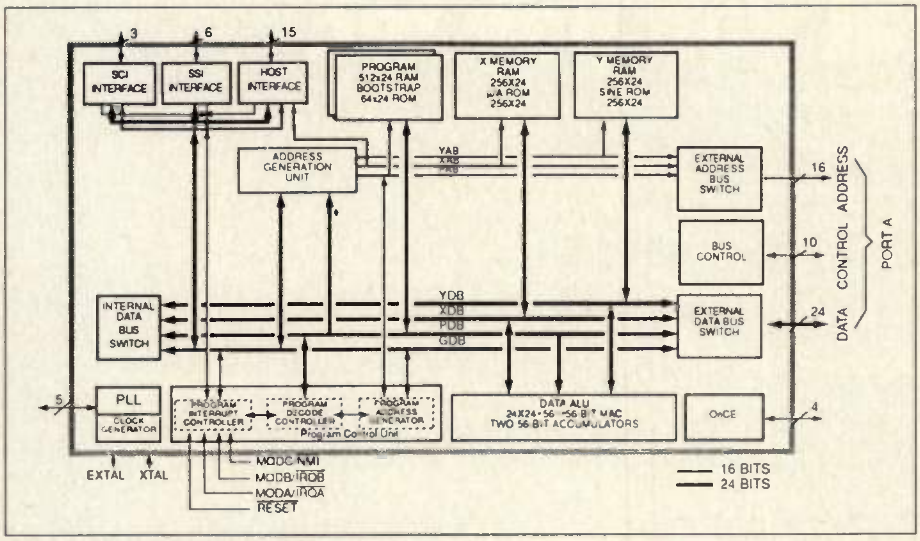

Figure 4. DSP56002 block diagram.

the L + R signal (4). The output of the I and Q modulators, of Figure 1 respectively are:

$$I = [L(t) + R(t)]\cos(\omega_c t) \tag{1}$$

and

$$Q = [L(t) - R(t)]\sin(\omega_c t) \tag{2}$$

These signals are summed and added to the carrier signal to result in

$$I + Q + \cos(\omega_c t) = [L(t) + R(t)]\cos(\omega_c t) + [L(t) - R(t)]\sin(\omega_c t) + \cos(\omega_c t) \tag{3}$$

which can be expressed as

$$B(t)\cos[\omega_c t + \gamma(t)] \tag{4}$$

where

$$B(t) = \sqrt{[1 + L(t) + R(t)]^2 + [L(t) - R(t)]^2} \tag{5}$$

and

$$\gamma(t) = \tan^{-1}\left(\frac{L(t) - R(t)}{1 + L(t) + R(t)}\right) \tag{6}$$

The limiter then serves the purpose of creating a constant amplitude signal while maintaining the unique phase information.

$$A\cos[\omega_c t + \gamma(t)] \tag{7}$$

This signal essentially becomes the carrier which is modulated by the L(t) + R(t) signal. The resulting signal from the transmitter is:

$$T_x(t) = A[1 + L(t) + R(t)]\cos[\omega_c t + \gamma(t)] \tag{8}$$

It is possible to receive this signal undistorted on a non-stereo system with an envelope detector. Because an envelope detector ignores the phase information contained in the signal (5), the result will be

$$A[1 + L(t) + R(t)]\cos(\omega_c t) \tag{9}$$

which is the standard signal which would be seen by the receiver if the transmitter was monaural.

A block diagram of the analog receiver is shown in Figure 2. The output of the envelope detector is simply the L + R + C(carrier) signal normally expected from a monaural system. Using this output and comparing it to the QAM output, the input to the QAM portion of the block diagram shown in Figure 2 can be gain modulated to remove the cosγ term (the input signal is divided by the cosγ term), with the resulting output of the I and Q demodulators being the required signals (3):

$$[1 + L(t) + R(t)] \tag{10}$$

and

$$[L(t) - R(t)] \tag{11}$$

The output of the QAM detector without gain modulation contains the L + R and the L − R encoded signals in the form given below:

$$I(t) = A[1 + L(t) + R(t)]\cos\gamma(t) \tag{12}$$

$$Q(t) = A[1 + L(t) + R(t)]\sin\gamma(t) = A[L(t) - R(t)]\cos\gamma(t) \tag{13}$$

The resulting signals after gain modulation are added together to produce the left only signal and subtracted to produce the right only signal.

Implementation Considerations

Figure 3 is a detailed diagram of the DSP based system. The input to the system comes from the AM receiver front end tuning circuitry and has been downconverted to the 450 kHz IF frequency. The sampling rate at the IF frequency would have to be at least 900 kHz to meet the Nyquist criteria (6). The AM signal is confined to a 10 kHz channel, however, so if the signal is further downconverted, a smaller sampling rate could be used to reconstruct the signal. As shown in Figure 3, the IF signal is downconverted to approximately 25 kHz and a sampling rate of four times that is employed in the sigma-delta A/D convertor (7). The analog multiplier shown in Figure 3 creates all ranges of harmonic frequency contents above the fundamental carrier frequency, However, sigma-delta A/D converter technology is based on oversampling and decimation processing technology, it eliminates the need to use an anti-aliasing filter (8,9).

The resulting digital signal now becomes the input to a general purpose digital signal processor. All functions to the right of the gray line are firmware solutions. The first function necessary is to separate the signal to retrieve the I and Q channels by demodulation to baseband. The digital input samples are multiplied by the current sine and cosine values of the numerically controlled oscillator (NCO). This operation is performed at the input sample rate of approximately 100 kHz. This is at a sample rate that is over four times the needed rate for reconstruction of the 10 kHz signal, therefore, the results of the last operation are low pass filtered and decimated to a sampling rate of approximately 25 kHz.

DSP Architecture

The Motorola DSP56002 was used for the digital radio firmware implementa-

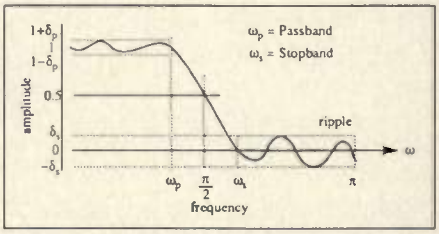

Figure 5. Design criteria for each stage half-band filter.

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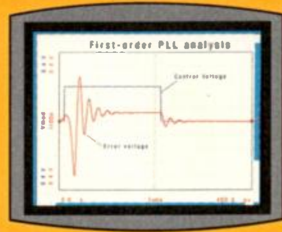

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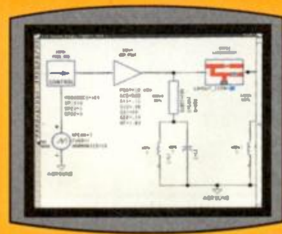

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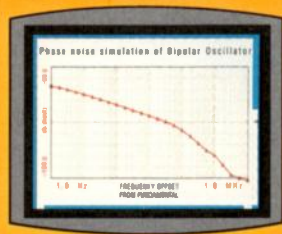

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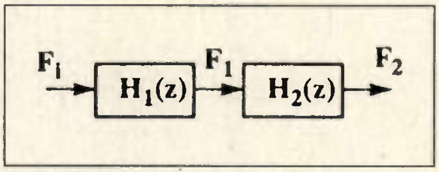


Figure 6. Two-stage cascaded half-band filter structure.

tion. The DSP56002, shown in Figure 4, is a fixed point processor with a dual Harvard architecture and a 56-bit accumulator for full 24-bit by 24-bit multiplies. The 56-bit accumulator is particularly well suited for audio applications due to the additional accuracy offered by the extended accumulator length. The DSP56002 has an on-chip PLL so that it can be operated at the fully specified processor speed by locking to, and multiplying up, a low speed external clock source. The ability to operate external to the DSP56002 at a lower speed reduces the signal interference problems in a design.

Halfband FIR Filter Design

Since the phase response is very important to decode the stereo signal and for the DPLL, a linear phase FIR filter is used which utilizes 4:1 decimation and LPF. Consider a special FIR filter called the half-band filter, which gives the required filter response and is computationally much simpler than conventional FIR filters. The half-band filter frequency response is symmetric about half the Nyquist frequency. This constraint results in half of the filter coefficients being exactly zero. Clearly if this symmetry can be tolerated in a design, considerable computational complexity can be avoided.

Let $H_h(z)$ be a linear-phase FIR filter transfer function with an odd number of coefficients, N. Since the linear-phase FIR filter has symmetric coefficient values, the frequency response $H_h(e^{j\omega})$ is real-valued function as

$$H_h(e^{j\omega}) = \sum_{n=0}^{M} b_n \cos(n\omega) \tag{14}$$

where M = (N-1)/2. Assuming the cutoff frequency of the half-band filter is $\omega_c = \pi/2$, we have

$$\omega_p + \omega_s = \pi \tag{15}$$

where ω_p and ω_s denote the passband and stopband edges, respectively. Moreover, assume that the passband ripple and the stopband ripple are the same (i.e., $\delta_p = \delta_s = \delta$). Thus, the response exhibits symmetry around $\pi/2$. Figure 5 shows a plot of such a symmetric halfband FIR filter response. In view of the symmetry, it is straightforward to show that

$$H_h\left[e^{j\omega}\right] = 1 - H_h\left[e^{j(\pi-\omega)}\right] \tag{16}$$

Also, in terms of H(z),

$$H_h(z) + H_h(-z) = 1 \tag{17}$$

Substituting equation 14 in equation 17, it can be shown that the coefficients, b_n, have the following constraints:

$$\begin{aligned} &b_{2n} = 0, \quad n \neq M \\ &b_M = h(M) = 0.5 \end{aligned} \tag{18}$$

As a result, the impulse response sequence has every odd-number sample equal to zero for M odd (except the coefficient $h(M) = b_M = 0.5$). Since the half-band filter is based on a symmetrical FIR design, the number of multiplications in implementing such a filter is one-fourth of that needed for arbitrary FIR filter designs.

The halfband filter coefficients can be obtained via the Fourier series method using a Kaiser window (10). By choosing proper parameters in the Kaiser window, more than –120 dB of stopband attenuation with specified transition bandwidth can be realized. Perhaps the only real drawback of the filter design is the requirement that $\delta_p = \delta_s$ shown in Figure 5. For most practical systems we have $\delta_s << \delta_p$. However, since the design curves are relatively insensitive to δ, the computational price paid for designing a filter with δ_p much less than required is generally small, and almost always much less than the 2:1 speedup achieved by these filters.

The halfband symmetric filter has a natural application in decimators with sampling rate changes of 2. However, since a halfband filter can only achieve a 2:1 decimation, a series of such filters should be cascaded to perform a higher decimation filter process. To obtain 4:1 decimation LPF with linear-phase response, a cascaded structure of two half-band filters has been implemented. Figure 6 shows the block diagram of a 2-stage decimation structure.

F_i in Figure 6 denotes the sampling rate of the input signal to the decimation filter. Note that the decimation of 4:1 can be achieved by two 2:1 decimators in cascade. For this implementation, an 11-tap FIR filter is used for the first

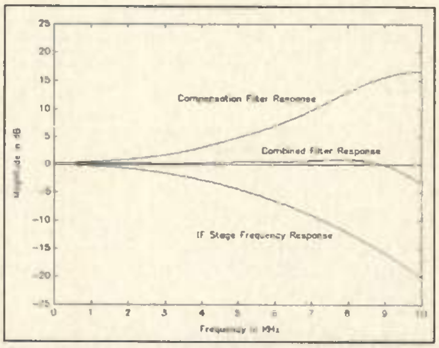


Figure 7. Frequency responses of the IF stage and compensated filter.

stage filter, while the second is done with 31 taps. This is efficient in reducing the number of operations performed per input since the signal is downsampled to approximately 50 kHz in the first filter using only 11 taps. The second filter, which takes 31 taps, is performed at the lower sampling rate.

For decimation, this cascaded half-band filter structure has the following advantages: significantly reduced number of computations; reduced memory requirement; simplified filter design problem; and reduced finite-word-length effects (11).

Let's examine the lowpass filter frequency regions for the individual stages. For the kth stage decimator, when k = 1, 2, the lowpass filter's passband, transition band and stopband regions are

$$\begin{aligned} &0 \le f \le F_k && \text{: kth stage passband} \\ &F_p \le f \le F_k - F_p && \text{: kth stage transition} \\ &F_k - F_p \le f \le F_k && \text{: kth stage stopband} \end{aligned} \tag{19}$$

where F_p and F_k are the passband frequency and the sampling frequency of the kth stage output, respectively. Signal energy in the transition band will alias back upon itself (after decimation by 2:1) only from F_p up to $F_{k-1}/2$; hence the baseband, $0 \le f \le F_p$, is protected against aliasing. The output of this decimation (D) filtering is

$$I_D(k) = A\left[1 + L(k) + R(k)\right]\cos(\gamma(k)) \tag{20}$$

$$Q_D(k) = A\left[1 + L(k) + R(k)\right]\sin(\gamma(k)) \tag{21}$$

where

$$\gamma(k) = \tan^{-1}\left[\frac{L(k) - R(k)}{1 + L(k) + R(k)}\right] \tag{22}$$

which is the discrete form of the continuous time equations 1, 2 and 6.

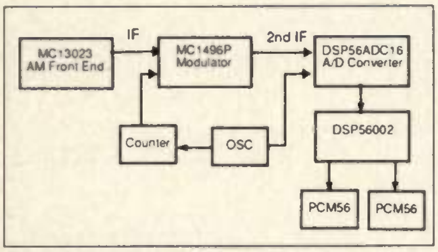


Figure 8. Block diagram of demo board.

Compensation Filter Design

Following the decimation filters are the filters which compensate the signal for the poor frequency response of the typical AM receiver front-end. A typical AM receiver front-end is analog and cuts the bandwidth to about 3.5 kHz as shown in Figure 7. This performance can be improved by applying a linear phase frequency compensation filter to the signal at the output of the 4:1 decimation filter stage.

Consider a filter which can compensate the frequency distortion by the front-end analog band-limiting filter. The objective is to design a compensation FIR filter whose transfer function $H_c(\omega)$ is such that its magnitude response approximates the inverse transfer function of the front end and has the linear phase property. Since the resulting FIR coefficients are real numbers, $H_c(\omega)$ is an even function in the interval $-\infty < \omega < \infty$. It is convenient to define the normalized frequency $0 \leq v \leq 1$ where $v = f/f_N$, so that the normalized Nyquist frequency $f_N = 1$. Using the Fourier series representation the transfer function can be expressed as (12)

$$H_c(v) = \sum_{n=-\infty}^{\infty} h_c(n)e^{jn\pi v} \tag{23}$$

where the Fourier series coefficients $h_c(n)$ are obtained by

$$h_c(k) = \frac{1}{2}\int_{-1}^{1} H_c(v)e^{-jn\pi v} \tag{24}$$

Since $H_c(v)$ is an even function in the interval of $-1 < v < 1$, equation 24 can be modified by

$$h_c(n) = \int_{0}^{1} H_c(v)e^{-jn\pi v}\cos(n\pi v)dv \quad n > 0 \tag{25}$$

and $h_c(-n) = h_c(n)$. Since the transfer function $H_c(v)$ is an unknown arbitrary function, it is extremely hard to find a closed form solution of the integration in equation 25. A numerical integration technique is used to find a set of coefficients for any shape of transfer function (13). It is important to divide the given interval into N smaller intervals (Δ), so that the discontinuity between subdivisions becomes minimum. The numerical approximation of equation 25 can be expressed as

$$h_c(n) \cong \sum_{i=1}^{N} \int_{v_{i-1}}^{v_i} H_c(v)e^{-jn\pi v}\cos(n\pi v) \tag{26}$$

Hence, approximating equation 25 by equation 26 amounts to finding the solution for the subinterval (v_{i-1}, v_i). There are numerous methods to find the solution (13), however, it has been found that Simpson's method gave us the optimum estimation on this implementation.

For convenience, let's define

$$g_i = \int_{v_{i-1}}^{v_i} H_c(v)e^{-jn\pi v}\cos(n\pi v)dv \tag{27}$$

Applying Simpson's rule on equation 27, the FIR filter coefficients $h_c(n)$ defined in equation 23 can be obtained as (13)

$$h_c(n) = \frac{\Delta}{6}\left[g_i + g_N + 2\sum_{i=1}^{N-1} g_i + 4\sum_{i=1}^{N-1} g_{i-1/2}\right] \tag{28}$$

However, it is not practical to have an infinite number of coefficients as shown in equation 23. So the series summation in equation 23 can be truncated at a predetermined finite number as

$$\bar{H}_c(v) = \sum_{n=-L}^{L} h_c(n)e^{jn\pi v} \tag{29}$$

The truncation in equation 29 can create the leakage phenomena which can be smoothed by a window function (10). In this implementation the Blackman-Harris window has been used to optimize for maximum side-lobe attenuation. Using the window function w(n), the FIR filter coefficients of equation 29 can be rewritten as

$$\bar{h}_c(n) = w(n)h_c(n) \quad \text{for} -L < n < L \tag{30}$$

where the Blackman-Harris window function is defined as (13)

$$w(n) = 0.358 + 0.488\cos\left[\frac{\pi n}{N}\right] + 0.1412\cos\left[\frac{2\pi n}{N}\right] + 0.011\cos\left[\frac{3\pi n}{N}\right] \tag{31}$$

Since the frequency response of the compensation filter requires only monotonic changes with respect to frequency, a simple (11-tap) symmetric FIR filter can compensate the frequency drooping to make a flat spectral response as shown in Figure 7. This effectively enhances the frequency response of the signal before further processing, and thus improves the quality of the final audio output. The typical frequency response, the compensation filter response and the combined filter response are shown in Figure 7.

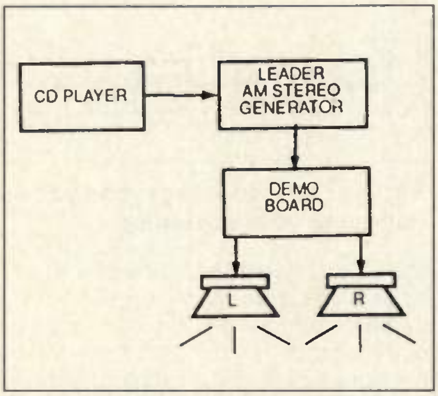


Figure 9. Test system block diagram.

Detection

Following the frequency compensation (FC) filtering, $I_{FC}(k)$ and $Q_{FC}(k)$ are the inputs to the digital equivalent of envelope detection in the first process of recovering the left and right signals. If the effects of phase error are ignored, then

$$I_{FC} \cong A\left[1 + L(k) + R(k)\right]\cos\left[\gamma(k)\right] \tag{32}$$

and

$$Q_{FC} \cong A\left[1 + L(k) + R(k)\right]\sin\left[\gamma(k)\right] \tag{33}$$

and H(k) is the output of the digital envelope detection:

$$\begin{aligned} H(k) &= \sqrt{I_{FC}^2(K) + Q_{FC}^2(k)} \\ &= A\left(1 + L(k) + R(k)\right)\sqrt{\sin\gamma^2 + \cos\gamma^2} \\ &= A\left(1 + L(k) + R(k)\right) \end{aligned} \tag{34}$$

H(k) can then be used to estimate $1/\cos(\gamma(k))$ where C(k) is this estimate.

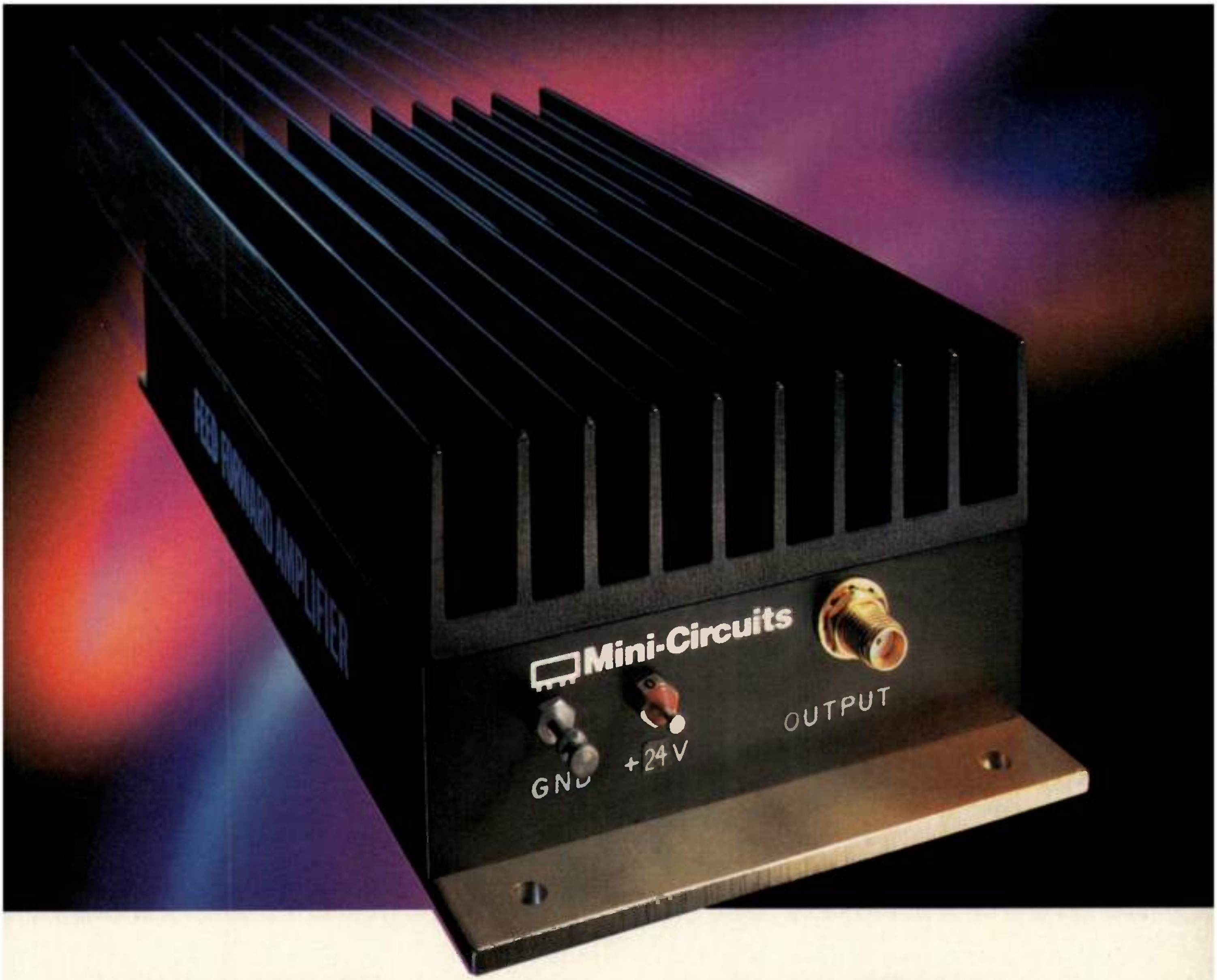
An error signal can then be formed since H(k) is known and since the estimate C(k) multiplied by $I_{FC}(k)$ should equal H(k), if the estimate has converged.

$$e(k) = I_{FC}(k)C(k-1) - H(k) \tag{35}$$

The error signal e(k) can then be used to update the estimate by

$$C(k) = \beta C(k-1) + (1-\beta)e(k) \tag{36}$$

β is selected based on the signal characteristics and it controls the rate of convergence of the estimate C(k). A value of 0.99 is typical, and is used in this application.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OUTPUT

At the time point, k, it is assumed that the iterative equation of 36 has converged and thus e(k) = 0 or, from equation 35:

$$C(k) = \frac{1}{\cos(\gamma(k))} \qquad (37)$$

Thus, I and Q signals can be modified by multiplying by this value of C(k) which results in:

$$\begin{aligned} I_E(k) &= I_{FC}(k)C(k) = A(1+L(k)+R(k)) \\ Q_E(k) &= Q_{FC}(k)C(k) = A(L(k)-R(k)) \end{aligned} \qquad (38)$$

which, with the exception of the carrier signal in the I channel, are the desired signals needed to obtain the left and right channel information. The carrier component is removed by a highpass filter. The same digital highpass filter is applied to both channels for two reasons. The first reason is that both channels need to have the same gain and amplitude characteristics to recover the right and left information correctly. The second reason is that the Q channel does have some phase information at baseband which needs to be removed so as not to distort the signal and will be discussed in the next section of this paper, but

$$\begin{aligned} I_{HPF}(k) &= A(L(k)+R(k)) \\ Q_{HPF}(k) &= A(L(k)-R(k)) \end{aligned} \qquad (39)$$

Now, by adding and subtracting the I and Q channels, the desired left and right channel signal can be found.

$$\begin{aligned} I_{HPF}(k)+Q_{HPF}(k) &= 2AL(k) \\ I_{HPF}(k)-Q_{HPF}(k) &= 2AR(k) \end{aligned} \qquad (40)$$

Digital Phase Lock Loop

All derivations to this point have ignored the phase error present when demodulating most real signals. A phase lock loop (PLL) is typically used to track and estimate the phase error. A digital phase lock loop (DPLL) was designed for the C-QUAM receiver and the following derivations include the phase error term. The input signal before demodulation looks like

$$R(k) = A[1+L(k)+R(k)] \cdot \cos\left(\frac{\omega_c}{\omega_s}(k) + \phi_e(k) + \gamma(k)\right) \qquad (41)$$

The numerically controlled oscillator (NCO) output would be:

$$N_I(k) = \cos\left(\frac{\omega_c}{\omega_s}(k) + \hat{\phi}_e(k)\right) \qquad (42)$$

where $\hat{\phi}_e$ is the current estimate of the phase error. The resulting demodulator output signal is

$$R_Q(k) = A[1+L(k)+R(k)] \cdot \sin\left(\gamma(k) + \left(\phi_e - \hat{\phi}_e\right)\right) \qquad (43)$$

plus double frequency terms which are filtered out by the low pass decimation filters previously discussed. After multiplying by the estimate of the inverse cosine term

$$A[1+L(k)+R(k)]\frac{\sin\left(\gamma(k) + \left(\phi_e - \hat{\phi}_e\right)\right)}{\cos\left(\gamma(k) + \left(\phi_e - \hat{\phi}_e\right)\right)} \qquad (44)$$

or

$$Q_E(k) = A[1+L(k)+R(k)] \cdot \tan\left(\gamma(k) + \left(\phi_e - \hat{\phi}_e\right)\right) \qquad (45)$$

It is straightforward to show that

$$Q_E(k) = A\left[\frac{(L(k)-R(k)) + (1+L(k)+R(k))\tan\left(\phi_e - \hat{\phi}_e\right)}{\left(1-\tan\gamma\tan\left(\tan\left(\phi - \hat{\phi}_e\right)\right)\right)}\right] \qquad (46)$$

Since $\tan(\phi_e - \hat{\phi}_e)$ is assumed small as $(\phi_e - \hat{\phi}_e)$ approaches zero

$$Q_E(k) = A\left[(L(k)-R(k)) + (1+L(k)+R(k))\tan\left(\phi - \hat{\phi}_e\right)\right] \qquad (47)$$

and using the same assumption for additional simplification:

$$Q_E(k) = A\left[(L(k)-R(k)) + \tan\left(\phi_e - \hat{\phi}_e\right)\right] \qquad (48)$$

The highpass filter discussed previously is now used to remove the baseband phase term and a lowpass filter is used to isolate this phase term and remove the left and right channel information:

$$T = \tan\left(\phi_e - \hat{\phi}_e\right) \qquad (49)$$

The new phase error estimate would typically be computed by:

$$\hat{\phi}_e(k-1) = \hat{\phi}_e(k) + \left(\phi_e - \hat{\phi}_e\right) \qquad (50)$$

However, since it is very processor inefficient on a DSP to perform inverse trigonometric functions, $\phi_e - \hat{\phi}_e$ is never calculated. The input to the loop filter is $\tan(\phi_e - \hat{\phi}_e)$ and the output is

$$T_{LF}(k) = \beta T_{LF}(k-1) + (1-\beta)T \qquad (51)$$

which is also in the form of $\tan(\phi_e - \hat{\phi}_e)$. Also note that if $x=\tan(\phi_e - \hat{\phi}_e)$, then

$$\sin\left(\phi_e - \hat{\phi}_e\right) = \frac{x}{\sqrt{1+x^2}} \qquad (52)$$

and

$$\cos\left(\phi_e - \hat{\phi}_e\right) = \frac{1}{\sqrt{1+x^2}} \qquad (53)$$

so that the sine and cosine of the new estimated phase error can be formed from trigonometric manipulation:

$$\sin\hat{\phi}_{e_{new}} = \cos\left(\phi_e - \hat{\phi}_e\right)\sin\hat{\phi}_{old} + \sin\left(\phi_e - \hat{\phi}_e\right)\cos\hat{\phi}_{old} \qquad (54)$$

and

$$\cos\hat{\phi}_{e_{new}} = \cos\left(\phi_e - \hat{\phi}_e\right)\cos\hat{\phi}_{old} + \sin\left(\phi_e - \hat{\phi}_e\right)\sin\hat{\phi}_{old} \qquad (55)$$

Since ω_c/ω_s = 1/4, only four values of sine and cosine need to be stored in a modular table to compute the NCO outputs which are then updated with the new phase error estimate using the same trigonometric manipulations used in equations 44 and 45. The final NCO output values are

$$N_I(k) = \cos\left(\frac{\omega_c}{\omega_s}(k) + \hat{\phi}_e\right) \qquad (56)$$

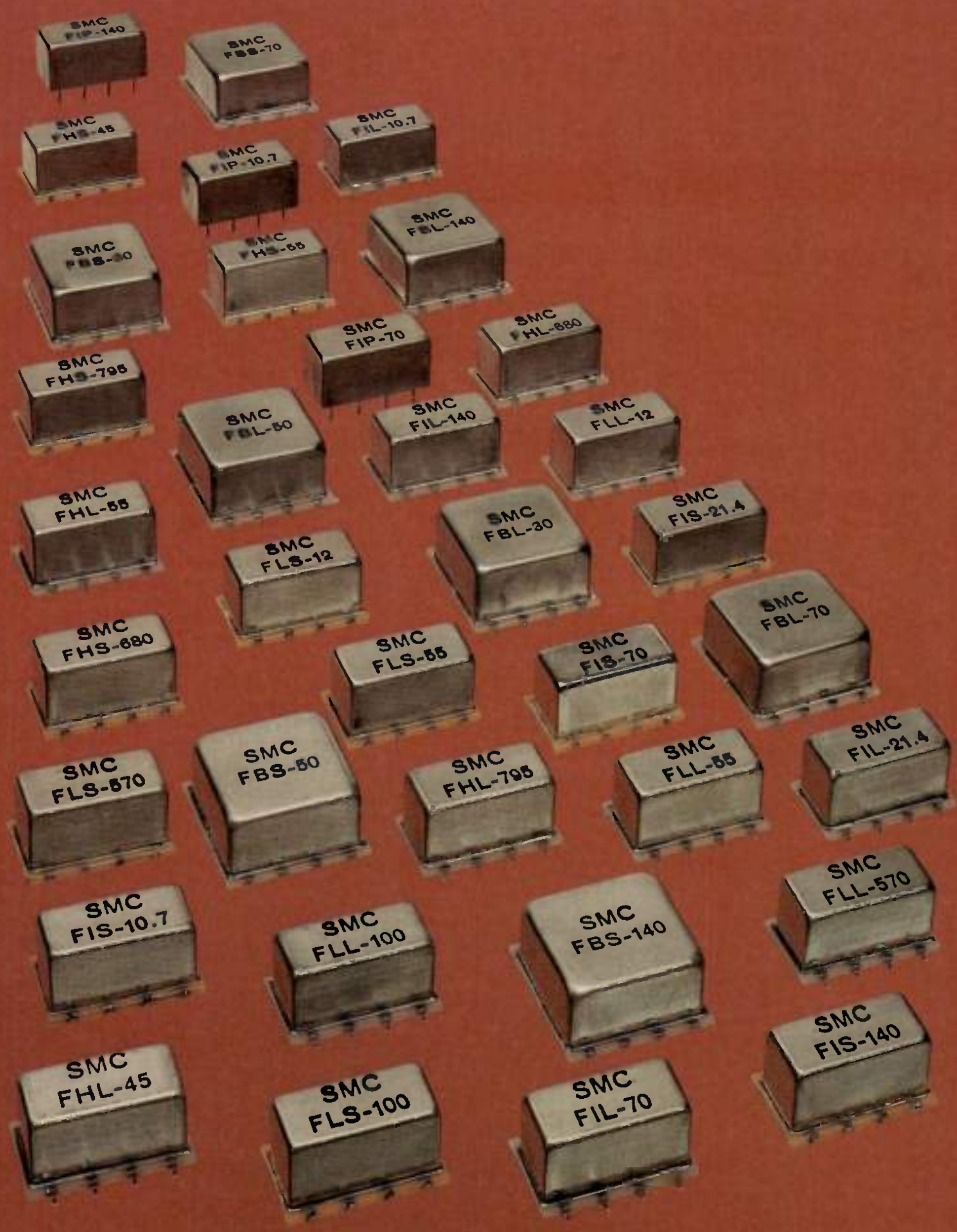
and

$$N_Q(k) = \sin\left(\frac{\omega_c}{\omega_s}(k) + \hat{\phi}_e\right) \qquad (57)$$

which can be calculated using trigonometric identities in terms of $\sin\hat{\phi}_e$ and $\cos\hat{\phi}_e$ as shown in the previous equations. Note that finding the value of $\hat{\phi}_e$ is never required using this expansion.

Hardware/Test

Figure 8 is a block diagram of the board built to test and verify the firmware solution. A Motorola MC13023 C-QUAM AM receiver front end is used to create the 450 kHz IF signal input to the Motorola MC1496 modulator. The output of the modulator is the input signal for the Motorola DSP56ADC16 sigma delta analog to digital convertor. The resulting digital output is then



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passed to the DSP56002 for processing. The Left and Right channel outputs of the DSP56002 are then converted to analog by the PCM56 and further amplified. Figure 9 shows the complete system used to demonstrate the board and the firmware.

Conclusion

The authors have shown that this combination of analog front end and firmware on a DSP can perform AM stereo reception. Additionally, the performance of this radio is much better than the fully analog counterpart, achieving more than 40 dB of channel separation.

Acknowledgments

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Rise Time/Fall Time Enhancement of Class C Bipolar Common Base Transistor Amplifiers

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

Class C RF defines a class of amplifiers which is non-linear, but highly power efficient, suitable for phase, frequency, or 100 percent pulse amplitude modulation. The circuit surrounding the amplifying device is simple compared to the linear class A and AB amplifier since no input bias supply is needed. At UHF and higher, grounded or common base is usually the preferred configuration because it offers higher gain than the grounded emitter case. A generic common base amplifier with shunt collector supply voltage feed is shown in Figure 1.

Generally speaking, for most applications the speed at which an amplifier of the type shown in Figure 1 responds to the RF drive is not too important. For example, when RF drive is applied to a class C RF power amplifier stage in a hand-held FM voice communication device, it is sufficient that the amplifier rise to its full output power before modulation is applied. Rise time may be as long as a microsecond, which is of no consequence in an application such as this. Also, when drive is taken away such as when reverting to the receive mode in a communicator, fall time may also be as long as a microsecond, which again, is of no consequence. Certainly, FM broadcast transmitters are not concerned about rise time or fall time because they are voice transmitters and essentially turn on or off only once per day or are continuously on.

Users of pulse modulation of class C amplifiers, such as that used in radar, with a pulse width as short as one microsecond, usually don't care about rise or fall time. If the ON time is considered to be that time between the 90 percent ON points, the RF drive pulse width may be adjusted or stretched to accommodate the slow class C amplifier.

There may be applications where the class C stage must respond to its RF drive pulse with far less rise time/fall

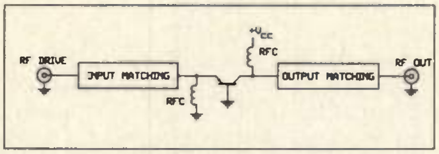

Figure 1. A typical class C common base RF amplifier topology.

time deterioration than that required by the applications just mentioned. As analog to digital converters gain in sampling rate and word size, wider bandwidth baseband signals (such as television video) can be accurately digitized for further use by the RF engineer.

It is possible to synthesize linear amplitude modulation onto a carrier by gating the RF carrier drive to a number of non-linear, but highly efficient RF sources. Each RF source must be of the correct power level, representing the weight of each digital bit, then power combined to a single output port. (1) The process and topology represent a high level digital to analog RF converter with carrier, or, a synthesized unbalanced mixer. This kind of modulation scheme requires that high level RF signal sources respond nearly instantaneously to the digital command to turn on or turn off compared to the shortest total on-time. The shortest total on or off time is equal to the period of the sampling frequency. In the case of digitized video (4.2 MHz bandwidth) where the sampling frequency must be at least 8.4 MHz, the on-time is as short as 119 nanoseconds. It is imperative that rise and fall times be only a small portion of this time or glitchy distortion is the result — that is to say, at least short enough so that the the glitches fall outside the RF passband and can be filtered out.

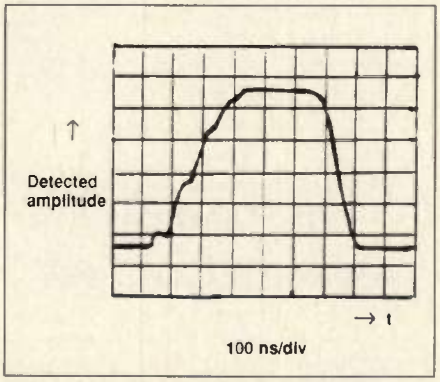

Figure 2. Rise/fall times of a typical 60 W, class C UHF bipolar power transistor driven with a drive pulse with 10 ns rise/fall times.

Figure 2 shows the detected RF output of a typical high power class C UHF transistor amplifier when driven with an assumed perfectly square drive pulse of RF (10ns rise/fall time).

Rise time deteriorates from 10ns to 300ns while fall time from 10ns to 125ns. The circuit used is a specific design derived from Figure 1 using a 60 watt device operating at 645 MHz. The device fails to turn on quickly because ON bias must be derived from drive. Until drive is applied, the transistor is biased off. Fall time suffers because stored charge in the device must bleed on its own since no external help is provided with circuitry. If a pulse of base-emitter current could be provided at the moment drive power is applied for the duration of the slow rise time, turn on time may be reduced to essentially that of the drive pulse. Likewise, if reverse

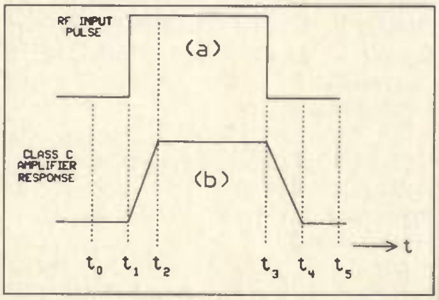

Figure 3. An assumed perfect RF drive pulse (a), and the detected class C output (b).

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	NEZ4450-8D, DD	9W	10.5	37%
	NEZ4450-4D, DD	4.5W	10.5	40%
5.9 to 6.4	NEZ5964-15D, DD	18W	9.0	33%
	NEZ5964-8D, DD	9W	9.5	35%
	NEZ5964-4D, DD	4.5W	10.0	37%
6.4 to 7.2	NEZ6472-15D, DD	18W	8.0	31%
	NEZ6472-8D, DD	9W	8.5	33%
	NEZ6472-4D, DD	4.5W	9.0	35%
7.1 to 7.7	NEZ7177-8D, DD	9W	8.0	31%
	NEZ7177-4D, DD	4.5W	8.5	33%
7.7 to 8.5	NEZ7785-8D, DD	9W	7.5	31%
	NEZ7785-4D, DD	4.5W	8.0	33%

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bias voltage is applied to the base-emitter junction at the moment that drive is removed, stored charge may be pulled out quickly reducing the fall time to essentially that of the drive pulse. It is the purpose of this presentation to show that external circuitry exists to perform these two functions and that it is automatic and self adjusting.

Slow Rise/Fall Time Analysis

If a gated RF drive envelope to a class C common base amplifier is assumed to be perfectly square at the leading and trailing edges, rise and fall time response of the class C stage will deteriorate. To be more specific, the output will rise to the top at a time later than that of the drive pulse and will return to zero at a time later than that of the drive pulse. Figure 3 illustrates this point graphically.

The rise time deterioration is t2-t1, while the fall time deterioration is t4-t3. The challenge is to reduce t2-t1, and t4-t3 to the smallest values possible. Furthermore, this must be done without ringing and overshoots. The only test equipment needed to measure the rise/fall time deterioration represented by Figure 2 are fast diode detectors, minimum stray capacitance on the lines carrying the detected signals, matching which minimizes reflections, a pair of directional couplers and a good dual trace oscilloscope.

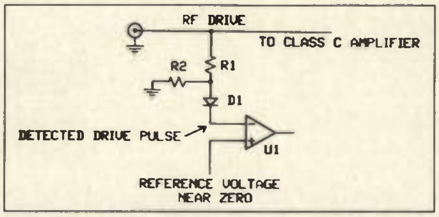

Figure 4. A circuit to detect the drive pulse and provide class A bias current to the transistor.

	t0	t1	t2	t3	t4	t5
Va	X	X	H	L	L	X
Vb	L	H	X	X	X	L
Vc	H	H	L	L	L	H
Vd	L	L	H	H	H	L
Ve	H	H	L	L	L	H

Table 1. A truth table is presented showing voltage values at various points in the circuit (see Figure 7).

A Solution

The following amplifier base bias conditions are sufficient to minimize t2-t1, and t4-t3.

• The transistor should be biased for class A operation during the rise time.

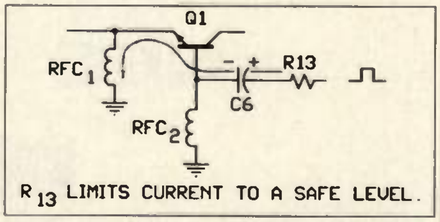

Figure 5. Scheme to bias Q1 class A during rise time.

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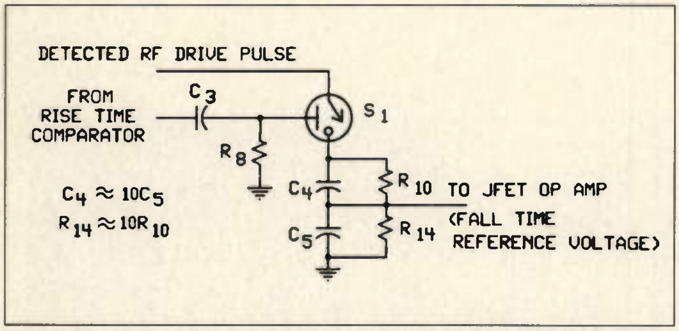

Figure 6. Circuit to sample and hold the fall time reference level.

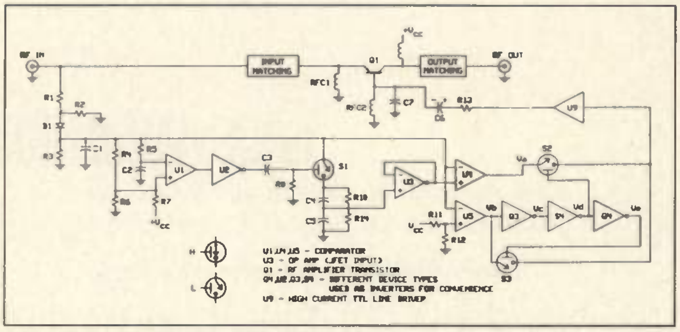

Figure 7. Representative circuit showing the functions to enhance rise/fall times.

• The transistor should be biased for class C operation after the rise time has expired and before the fall time is initiated.

• The amplifier should be biased for cut-off during the fall time.

In order for class A bias to be applied during the rise time, it is necessary to sense the presence of the drive pulse. A fast diode detector can be used for this purpose, causing a voltage comparator to change state. A functional circuit could be that of Figure 4.

The base of the class C amplifier could be AC coupled to the output of the comparator of Figure 4 to cause base-emitter current to flow during the rise time, biasing the stage in class A. Current DC is limited by means of a series resistor with the DC blocking capacitor. The capacitor is charged at the end of the rise time, reverting amplifier bias back to class C. A functional diagram is shown in Figure 5.

Since RFC2 is much larger than RFC1, the shot of base-emitter current largely bypasses RFC2 instead finding ground through RFC1 until the blocking capacitor is charged.

While the drive pulse remains on after rise time, the RF transistor is biased for class C. At fall time, another voltage comparator may be used to sense when the drive pulse envelope is falling and can change state early in the fall time. This changing state can be used to ground the positively charged side of the blocking capacitor in Figure 5. This causes an instantaneous reverse bias of

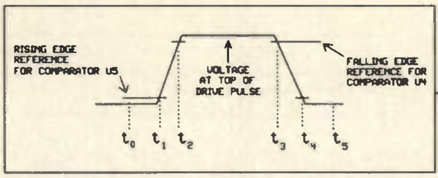

Figure 8. Rise time reference and fall time reference superimposed on detected drive pulse.

the base-emitter junction of the transistor, causing it to cut-off.

The previous three step process is easy to see when presented functionally. Even the reference for the rise time comparator is easy to establish since it must only be arbitrarily close to ground. The fall time reference must be self adjusting, however, since the top of the drive pulse envelope may be any detected voltage level depending on the RF drive power of the amplifier. It is sufficient to say that a reference for the fall time comparator must be derived from the detected drive pulse itself, then set somewhat below the top of the drive pulse envelope. The comparator will change state as the detected drive pulse falls below the reference. The reference can be established by means of a sample and hold circuit, and charge can be stored by a capacitive voltage divider driving a JFET input operational amplifier used as a voltage follower/buffer. Figure 6 shows the functional solution to the fall time reference circuit. The switch, one-fourth of a quad bi-lateral CMOS analog switch, is closed momentarily by AC coupling its gate to the rise time comparator output. When the switch opens, the peak detected drive pulse amplitude is stored across the two series capacitors. The ratio of the two capacitors determines the reference to peak drive voltage ratio.

For the composite circuit, it is only necessary to combine Figures 4, 5 and 6 into a practical configuration, deciding on component values and preventing logic impossibilities and short circuits from occurring. A complete functional diagram is shown in Figure 7.

A truth table (Table 1) describes the logic states leading up to establishing

- class A bias during rise time
- class C bias between rise time and fall time
- cut-off bias during fall time

Figure 8 indicates the position of the drive pulse and reference voltages at selected times corresponding to the logic level states of Table 1.

A realizable circuit with proven components is shown in Figure 9 (see page 50). Maximum economy of quad packages is used where one bilateral CMOS switch in a package of four is left over, so it is used as a logic inverter. Fast switching transistors and fast voltage

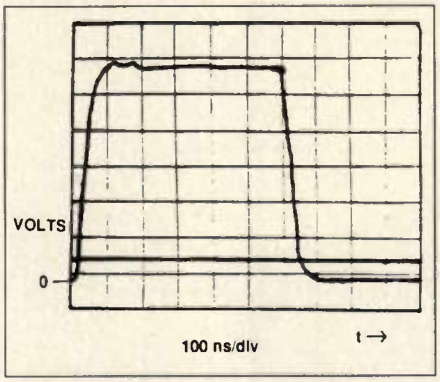

Figure 10. Detected output of enhanced amplifier along with rise time reference. Compare to Figure 2.

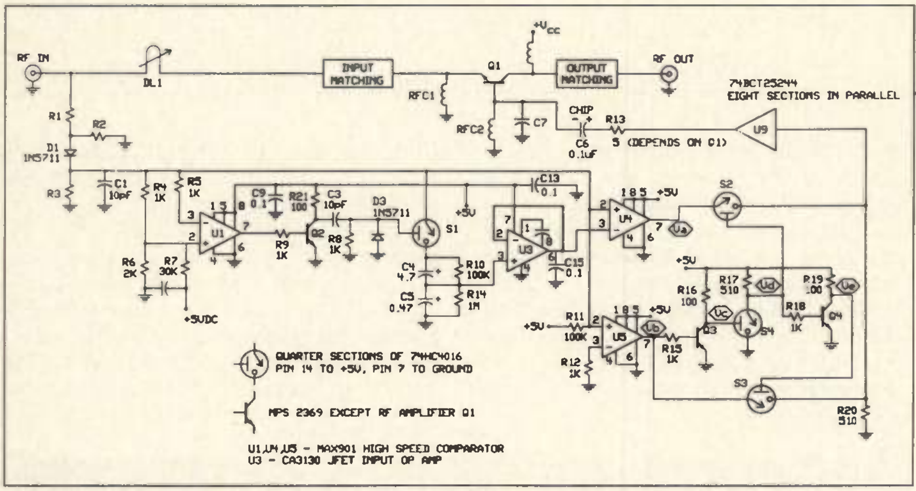

Figure 9. A practical rise/fall time enhancement circuit.

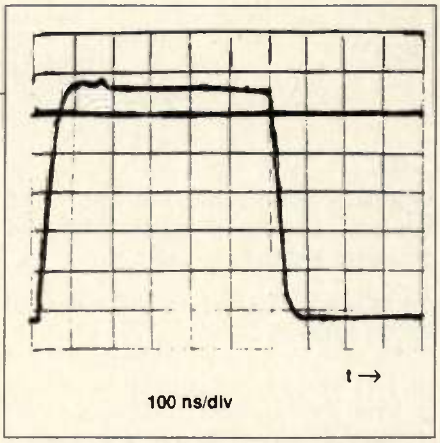

Figure 11. Detected output of enhanced amplifier along with fall time reference. Compare to Figure 2.

comparators are used throughout. JFET input voltage follower operational amplifiers are used where it is important to have very high input impedance so as not to discharge the hold capacitors. Circuit component values are real in a proven circuit for a 60 watt class C amplifier. RF components are not specified since they are unique to the design, but RFC2 must be five to ten times greater than RFC1 for the base-emitter junction to be pulsed properly. The fall time reference capacitors discussed and shown in Figure 6 are C4 and C5 in Figure 9. They are in a 10:1 ratio so that the fall time reference is set at 10/11 of the peak detected drive pulse value, or 91 percent.

Results

Figure 2 shows the typical grounded base response time of the unmodified class C amplifier.

Figure 10 shows the actual enhanced rise time (steepness of leading edge slope). The horizontal trace at the bottom is the rise time reference set by R6R7 in Figure 9. Likewise Figure 11 shows the actual enhanced fall time with the horizontal trace indicating the position of the fall time reference established across C4 and C5 in Figure 9.

Figure 12a and 12b show side by side the unenhanced output of the class C amplifier along with the enhanced.

It is felt that the circuit component values of Figure 9 are universal for any class C solid-state amplifier except those RF components unique to the

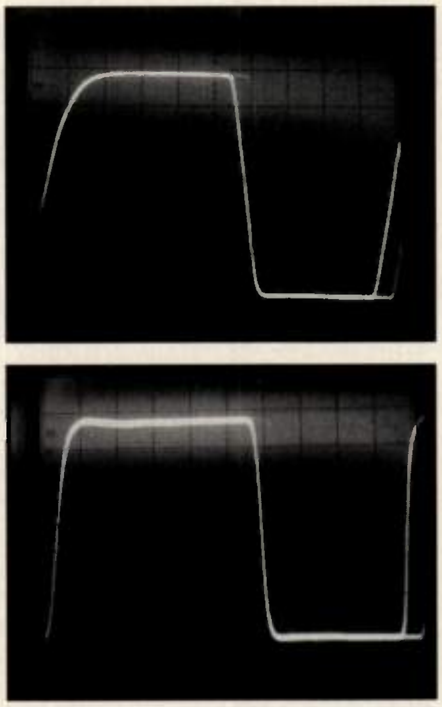

Figure 12. a) Unenhanced output pulse retuned to take on a better shape than that of Figure 2. b) Output pulse with rise time/fall time enhancement engaged.

amplifier itself and the value of C6 and R21. R21 is chosen to limit the base pulse current amplitude to a safe level for the RF transistor while C6 must be chosen to allow the base pulse to continue through the rise time and not to discharge completely during fall time until fall time is complete. This circuit has been tried successfully using RF transistors capable of 10 watts through 60 watts with changes only to these two component values.

A patent has been applied for for the Rise Time/Fall Time Enhancement method just described. For its commercial application, please consult the author. ***RF***

Reference

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About the Author

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New VCO Family is Created From Theory and Modeling

By Michael S. Vogas
Synergy Microwave Corp.

Synergy Microwave Corporation introduces a series of Voltage Controlled Oscillators (VCOs) covering the 400 to 1800 MHz range. These VCOs are extremely rugged, hermetically sealed and are offered in Synergy's 16 pin metal package, plug in, and surface mount form. These devices exhibit extremely low phase noise and are constructed using a minimum number of parts, making them a low cost design alternative to other VCOs. In this article, a theoretical/CAD oscillator design approach is presented that led to the design of this new product line.

Although numerous researchers in the past have published oscillator configurations based on some form of small signal analysis, these approaches require a combination of theoretical and empirical calculations and do not always lend themselves to the desired result as tuning of the final device is inevitable. Synergy's approach involves simulation with Compact Software's Harmonica Linear/Nonlinear Harmonic Simulator, resulting in considerable savings in design time as well as tuning time, yielding results very close to the desired performance.

Transistor selection is of primary importance. For a frequency range of a few hundred megahertz to several thousand megahertz, VCOs are commonly built around Bipolar Junction Transistors (BJTs) due to their inherent low noise. Contrary to popular belief, f_{max} is not the most definitive quantity in the selection of the device. f_{max} is defined as the frequency at which the unilateral gain falls to unity, and the device ceases to oscillate. With a high f_{max}, the device is susceptible to oscillations outside the desired frequency range, and extraneous coupling between the device leads may arise. Needless to say, if cost is to be kept low, devices with very high f_{max} are prohibitive.

f_{max} should definitely be within the desired range of operation, but power dissipation is also an important criteria in the selection of the device. Power dissipation is important, as VCO active devices operate at high power levels. Also, care should be taken to control the oscillation totally through the external circuitry with no possible spurious oscillations inside the device. A preliminary simulation/construction of a non-oscillating circuit is necessary in order to prove that the device is free of unwanted spurious oscillations. The line of BJTs selected offers an ideal choice as they can typically be used up to 8 GHz and lend themselves to all forms of external positive feedback, while producing an average negative resistance of 50 to 400 ohms (1).

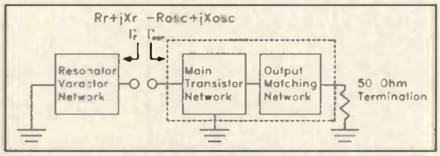

Figure 1. Block diagram of a typical oscillator network as it is used for graphical/computer analysis.

Basic Theory of Operation

Most of the available literature on oscillators describes the operation of oscillators using small signal S-parameters. The main reason for this occurrence is that small signal S-parameters are readily measurable and all transistor manufacturers supply them for their

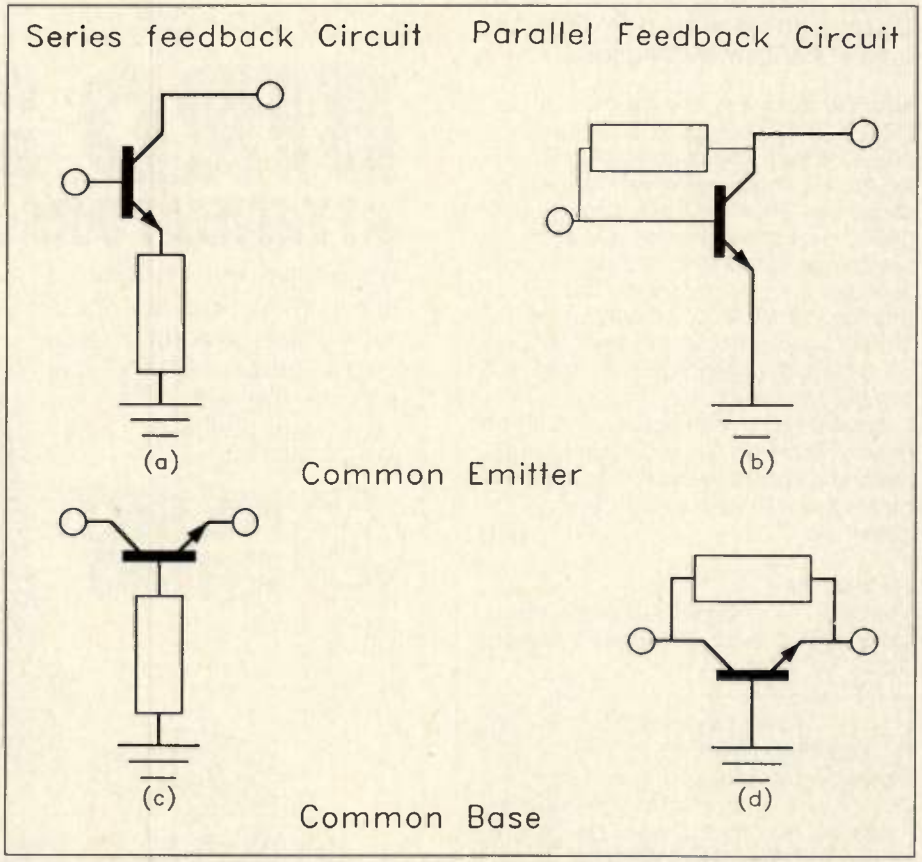

Figure 2. Typical BJT configurations for the main transistor network of Figure 1.

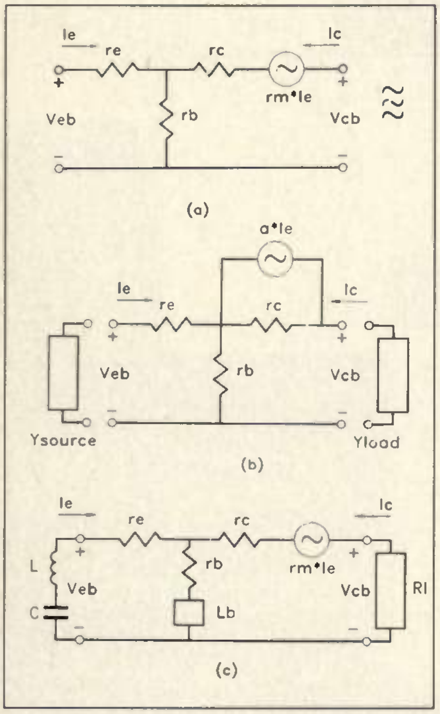

Figure 3. Small signal models for the configuration used.

devices. The small signal design technique offers valuable insight into the conditions that govern oscillator starting, while giving a rough prediction of the possible frequency range of operation. The disadvantage, though, is that those estimates are valid only for the first few cycles after the device is turned on.

A proper way to describe the oscillation process is through a thorough derivation of the feedback network of the device, with an application of the Barkhausen criteria. The Barkhausen criteria state that the product of the forward gain path and the feedback path for the complete oscillator network should be equal to unity, while the feedback loop should add the output wave form to the input of the combined network with no phase shift (i.e. positive feedback). During that condition the mechanism that keeps the device in oscillation is the continuous balancing of these two quantities. For example, as the oscillation amplitude of the device increases, its gain decreases to satisfy the previous criteria.

The derivation of the equations for the feedback loop and the forward gain of the device can be very complex in certain cases, and special care should be taken in identifying the two loops. Rohde (2) offers a very good description of that process. In addition to the above, a rather simplistic method of describing the oscillation criteria has been used in the industry. That process is not favored for its insight in the operation of the device, but rather for its ease of applica-

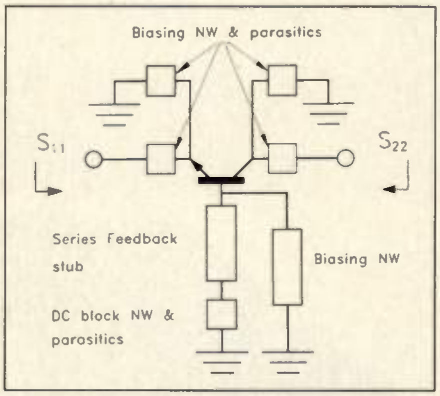

Figure 4. Transistor in CB showing biasing network and biasing parasitics.

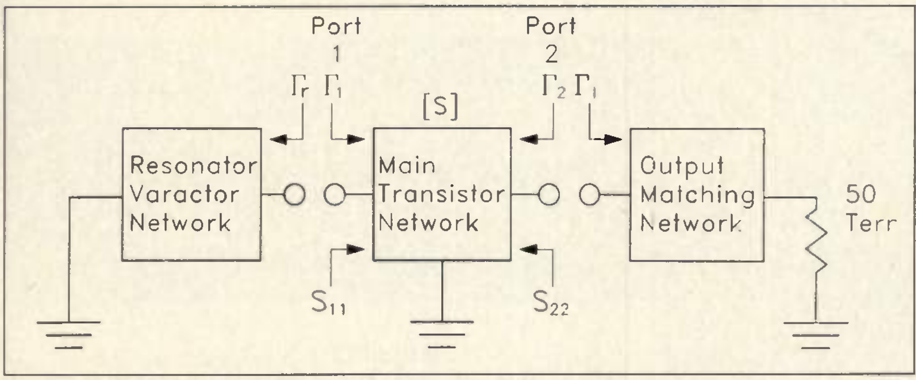

Figure 5. Block diagram at a typical oscillator network.

tion. In that latter approach, considerations for the input resistance of the device attempt to explain why the amplitude of the signal starts increasing while building up from the internal noise level. Looking at Figure 1, it can be seen that this happens because the resistance of the combined active device is greater than the resistance of the load, namely:

$$|-R_{osc}| > R_r \text{ and } X_{osc} = -X_r \quad (1)$$

As the oscillation amplitude keeps increasing, the DC operating point of the device shifts continuously thereby changing the g_m of the device and, of course, its input and output impedance. This situation produces a frequency shift, a change in output harmonics and/or a change in output power levels, and the oscillator operation predictions in general. The increase in amplitude will cease when the real parts of the respective impedance and the active device become equal, or:

$$|-R_{osc}| = R_r \text{ and } X_{osc} = -X_r \text{ or} \\ \text{alternatively when } \Gamma_r \cdot \Gamma_{osc} = 1 \quad (2)$$

Where Γ_r and Γ_{osc} are the reflection coefficients produced by R_r+jX_r and $R_{osc}+jX_{osc}$ respectively. When steady state occurs, the device reaches saturation levels. R_{osc} is a very important quantity and is accurately described by the nonlinear I-V relationship of the transistor. If we look at a common base transistor we can see that, due to the nonlinear current source, the value of R_{osc} varies non-linearly about a given I-V operating point. Since that resistance is related to the internal current source, we can imagine the total current as the sum of all the individual currents at all harmonic frequencies. This can be simply stated as:

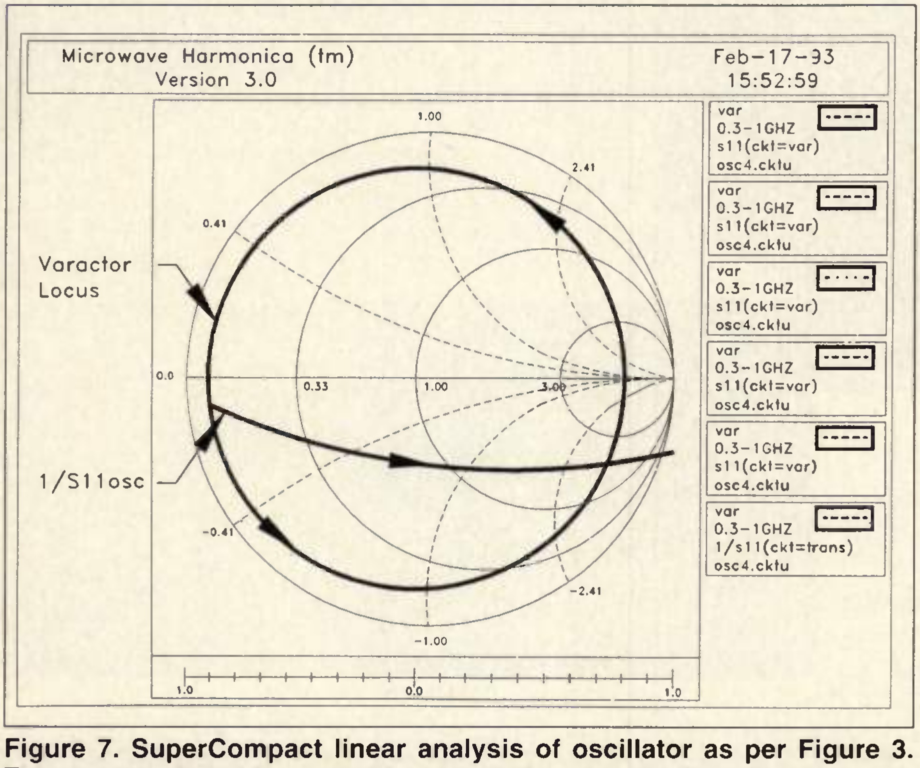

Figure 7. SuperCompact linear analysis of oscillator as per Figure 3. The varactor locus is created by cumulative varactor iterations for various capacitances.

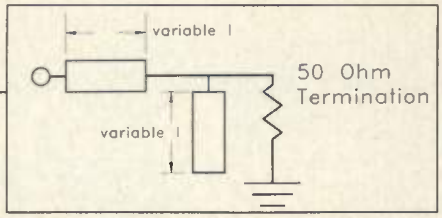

Figure 6. Typical output matching network showing variable line lengths used for optimization.

$$I_{total} = I_{dc} + I_1 + I_2 + I_3 + \ldots = \sum_{n=0}^{\infty} I_n \quad (3)$$

Where I_{dc} is the DC current through the device when it does not oscillate, and I_1, I_2, I_3 are the currents of the first, second and third harmonics, respectively, which come into play when the device starts oscillating. The total current mentioned before is related to the internal current generator of the transistor, and we can thus express the total transconductance characteristic of the device as:

$$G_{mtotal} = g_{m0} + g_{m1} + g_{m2} + g_{m3} + \ldots = \sum_{n=0}^{\infty} g_{mn} \quad (4)$$

In the above, g_{m0} is the value of transconductance at DC while g_{m1}, g_{m2}, etc. are the values at the first, second and third harmonics respectively. It is important to keep in mind that g_{m1}, the transconductance attributed to the first harmonic is approximately equal to $0.7g_{m0}$, and is the most important of all the harmonics in equation 4. As the oscillations increase in amplitude the value of g_{m1} decreases in order to keep the device meeting the oscillation criteria, and have G_{mtotal} equal to a constant value for a specific DC input current. The latter explains why the magnitude of R_{osc} is more negative while the system commences oscillation and settles down to a more positive number when the system reaches equilibrium. Also, the fact that the g_m terms are a decaying series verifies the fact that the system can not oscillate to infinity.

There are several options (3) available for the main transistor network, with some presented in Figure 2. A common base or common emitter configuration is used, with the feedback element in series or parallel form. The configuration selected for Synergy's designs is the common base (Figure 2c) because it lends itself more easily to oscillation and presents an inherent isolation from output (collector) to input (emitter). In this configuration we can assume that the internal AC base resistance r_b is much less than r_c, the internal AC collector resistance, an assumption which is true because r_b is less than 20 ohms while r_c, may well be in the ten Megohm range. Based on the above assumption, we

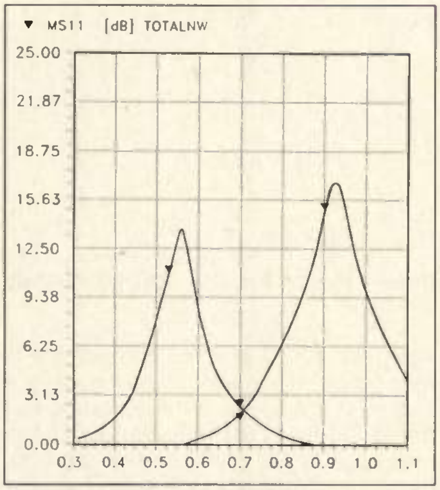

Figure 8a. Magnitude of output S11.

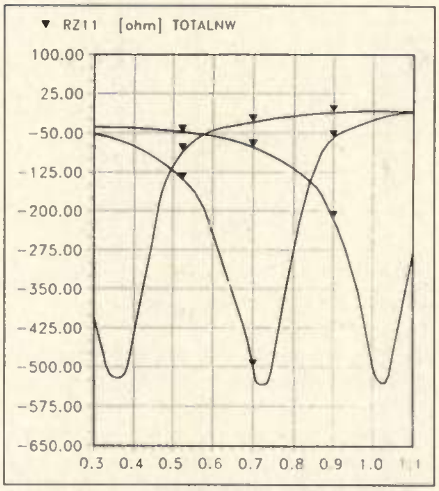

Figure 8b. Typical negative resistance input for the high, mid and low bands.

can safely assume that there is a factor α such that: r_m (transfer resistance of the BJT current generator) $\approx \alpha \cdot r_c$, where α is called the short circuit current multiplication factor, a quantity greater than or equal to unity. The importance of α in the determination of the oscillation conditions is easily seen if we attempt to find the input impedance of the previous circuit. It can be easily proven that the input impedance is:

$$R_{in} = \frac{V_{eb}}{I_e}\Bigg|_{I_e = 0} = r_e + \frac{r_b r_c(1-\alpha)}{r_b + r_c} \qquad (5)$$

$$\approx r_e + r_b(1-\alpha)$$

From 4 and 5 we can see that it is possible for a common base configuration to develop a negative input resistance (since $\alpha > 1$) if r_b (or r_b in series with another impedance) becomes sufficiently large. The latter can be easily satisfied by the addition of an external element like the microstrip inductance L_e (Figure 3). In order to examine the stability of such a network let us assume that the admittance matrix of the combined device and base stub is Y. In that case we have:

$$\begin{bmatrix} y_{11} & y_{12} \\ y_{21} & y_{22} \end{bmatrix} = Y \qquad (6)$$

If the output port is terminated with a frequency varying network Y_{load}, the input admittance is given by equation 7. Similarly if the input port is terminated by a frequency varying network Y_{source}, then the output admittance will be given by 8:

$$Y_{in} = y_{11} - \frac{y_{12}y_{21}}{y_{load} + y_{22}} \qquad (7)$$

$$Y_{out} = y_{22} - \frac{y_{12}y_{21}}{y_{source} + y_{11}} \qquad (8)$$

In view of equations 7 and 8, the stability of the device can be ascertained in two ways: the Linvill stability factor (equation 9), or the condition in equation 10 which states that the Z (or Y) matrix of the complete network should be equal to 0.0.

$$c = \frac{|y_{21}y_{12}|}{2\mathrm{Re}\{y_{11}\}\mathrm{Re}\{y_{22}\} - \mathrm{Re}\{y_{21}y_{12}\}} \qquad (9)$$

where $c > 1$ for instability

or [V] = [Z] · [I] where the determinant of [Z] is equal to 0.0 for the onset of oscillation. We shall use the second method which gives us a little more insight into the operation of the circuit. If we add the output impedance R_l, the

F(GHZ)	$\lvert S11_r \rvert$	$\lvert S11_{osc} \rvert$	$<S11_r$	$<S11_{osc}$
0.570	0.915	1.118	+178.8	−179.0
0.740	0.925	1.993	+150.7	−153.0
0.930	0.930	1.175	+156.0	−156.0

Notice that $|S11_r| \geq \frac{1}{|S11_{osc}|}$

and $< S11_r = - < S11_{osc}$

Table 1. Table of the elements of Figure 3 as the varactor is tuned across the band.

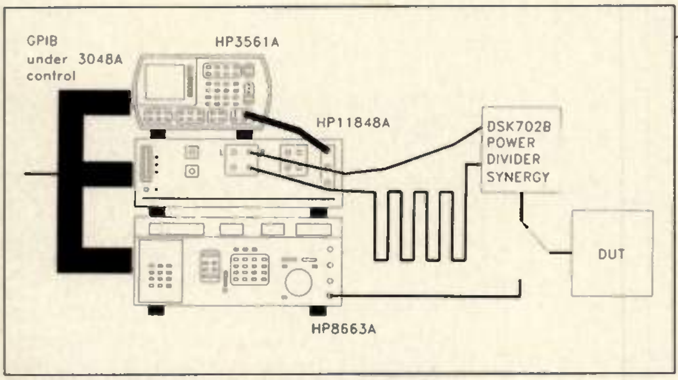

Figure 9. Typical phase noise setup.

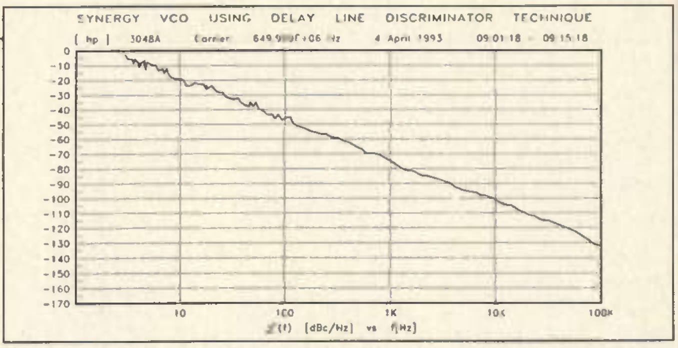

Figure 10. Typical phase noise for the 500-1000 MHz unit.

microstrip inductance L_b and the varactor/inductor combination to the network we get a network similar to the one in Figure 3c. The following loop equations can then be derived:

$$\begin{bmatrix} 0 \\ 0 \end{bmatrix} = \begin{bmatrix} j(\omega L - 1/\omega C) + r_e + r_b + L_b & r_b + L_b \\ r_m + r_b + L_b & r_c + R_r + r_b + L_b \end{bmatrix} \begin{bmatrix} I_e \\ I_c \end{bmatrix} \tag{10}$$

The determinant of the above matrix must reduce to 0, so the above equation becomes:

$$j(\omega L - 1/\omega C)(r_c + R_1 + r_b + L_b) + r_e \cdot (r_c + R_1) + (r_e + R_1 + (1-\alpha)\cdot r_c)\cdot (r_b + L_b) = 0 \tag{11}$$

By setting the reactive part to 0.0 we can show that the resonant frequency is:

$$fo = (1/2\pi)\cdot\sqrt{(1/LC)} \tag{12}$$

Similarly, by setting the real part to 0 we can show that:

$$\alpha \geq \left[1 + \frac{R_l}{r_c}\right]\cdot\left[1 + \frac{r_e}{r_b + L_b}\right] \tag{13}$$

If $\alpha \geq 1$ in 13, by properly choosing R_l (i.e. the output matching network) and L_b (4) (the length of the microstrip inductor), we can ascertain that the circuit will oscillate. Furthermore, by properly selecting the value of the capacitance/inductance at the emitter, we can tune the circuit over a range.

Circuit Simulation

A proper method for simulation of oscillating circuits is the nonlinear approach, which can be achieved by using Microwave Harmonica Ver. 3.0 and its harmonic balance techniques. The approach describes the active device with a nonlinear large signal S-parameter model, and simulates the network using well defined goals for oscillation ranges, fundamental and output harmonic power levels. The linearized approach described in the following section (5) involves a simplistic method for the design of the unit. It offers a prediction of the output oscillation range but it has no insight into any of the circuit performance parameters associated with the nonlinear parts of the models.

The first step in the simulation is to determine the optimum base reactance at the desired frequency range of operation (Figure 4). This reactance is selected to make certain that the transistor will oscillate at one frequency in the region of interest (preferably mid-band). The value of that inductance can be arrived at by a simulation of the transistor in CB configuration with the base stub present and including all parasitics, DC block capacitors and biasing networks. Such a network is shown in Figure 4. The best choice for a stub length is one that achieves maximum amplitude for S11 and S22. In general, the longer the stub, the lower the frequency and vice versa (4). The Optimization Option in SuperCompact is ideal for arriving at the correct value of the base inductance. Care should be taken to see that in the process of maximizing S11 and S22, S21 is kept relatively constant as that quantity will now roughly govern the linearity of the output power. To establish the conditions of oscillation in Figure 5, the following have to be met:

$$\Gamma_r \cdot S_{11} = 1, \quad < \Gamma_r \leq S_{11},$$

$$\text{and } \Gamma_1 - S_{22} = 1 < \Gamma_1 \leq S_{22}$$

We can continue the optimization by matching the output of the transistor, S22, to the 50 ohm load (Figure 5). This is usually achieved by maximizing S11 while the output is conjugately matched. A typical output matching network is a combination series/open stub (Figure 6). During the optimization process in

Range	Nominal Power Output	Flatness	2nd Harmonic
(MHz)	(dBm)	(dB)	(dBc)
400-800	14.5	3	9
500-1000	14	3	10
600-1200	14	3	10
700-1400	14.5	3	10
800-1600	13.5	3	10

Table 2. Device specifications.

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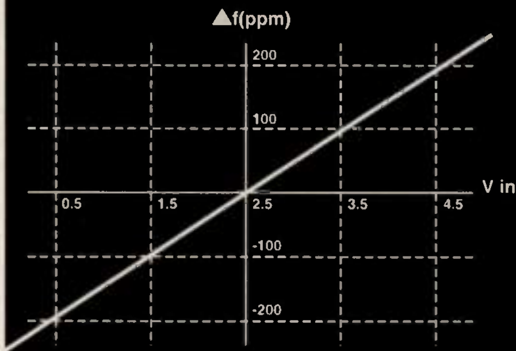

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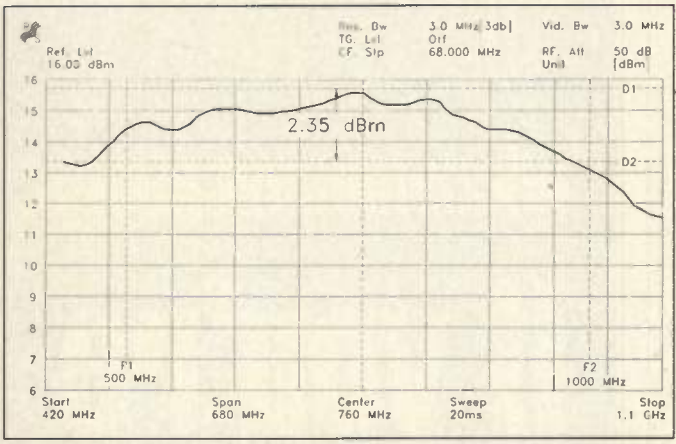

Figure 11. Typical power output for the 500-1000 MHz unit.

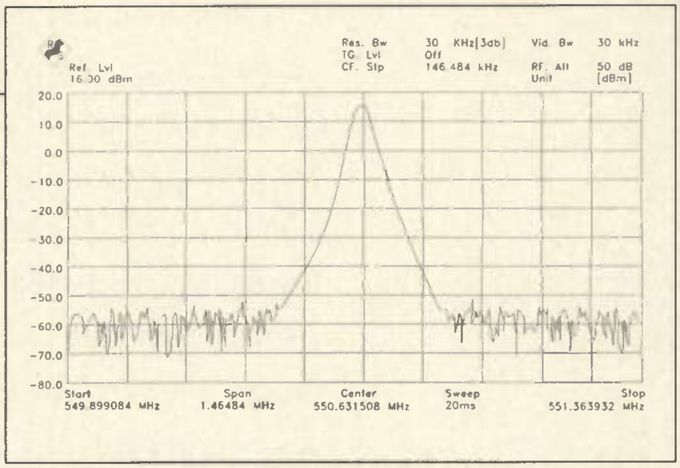

Figure 13. Typical lobe power output.

Microwave Harmonica, the lengths of the series and open stubs are left to vary freely while a goal of matching a 50 ohm load has been selected. Of course, this particular matching network is not the sole solution, just a simple choice out of many (some very complex) that the designer can use for effective matching to the load. Once the output has been matched, the input of the one port network, which now looks similar to the combined network in Figure 1, is plotted. If the matching at the output was performed properly, the value of its reflection coefficient should still be much greater than one. The output can either be plotted as $1/S11_{osc}$ on the Smith chart in order to match any loops in the locus, or in a tabulated form in order to match amplitude and phase to the reflection coefficient of the resonator. Loops in the locus of the active device at this point may indicate a possibility for injection locking or spurious oscillations (see Figure 7).

The resonator section is usually composed of a network of varactors, with a short-circuited stub to ground which serves as a matching network to the input of the transistor combination. Both sections of the circuit as per Figure 1 were simulated simultaneously using Microwave Harmonica, with the results shown in Figure 7. The varactor locus is achieved by using successive sweeps while changing the varactor capacitance. A tabulated form of the results is shown in Table 1 clearly showing the oscillation condition for 500-1000 MHz VCO.

The output value of S11 for the combined oscillator network is shown in Figure 8a for two values of the varactor capacitances one at the low and the other at the high end of the 500-1000 MHz VCO. Note that the output amplitude variation is minimal and the output magnitude of S11 is better than 13 showing a robust oscillation across the range. For the same system the value of the output impedance, approximately -500 ohms, is very constant across the range and is seen in Figure 8b.

The devices exhibit excellent phase noise characteristics as seen in the previous two figures. The phase noise was measured using the phase discriminator method using an HP 11804 phase noise measurement system. A typical layout is shown in Figure 9 and a typical phase noise plot is shown in Figure 10. The noise figure is –100 dBc at 10 kHz from the carrier and –75 dBc at 1 kHz.

The output power variation and the tuning voltage across the range for a 500-1000 MHz VCO are shown in Figures 11 and 12 respectively. A typical oscillator lobe is shown in Figure 13. The VSWR is measured to be 1.85:1 across the band or better. The tuning sensitivity is 50-70 MHz/V and the absolute maximum tuning voltage is 25 volts. Typical frequency pulling is 17 MHz peak to peak and frequency pushing 3 MHz per volt.

A table of the specifications of the five devices that cover the above mentioned range is shown in Table 2. This represents only the initial range to be introduced; a complete family of devices covering all commercial ranges is in development.

Conclusion

A concise description of a method for the design of a VCO has been presented, and the results of applying this method to five new VCO models have been demonstrated.

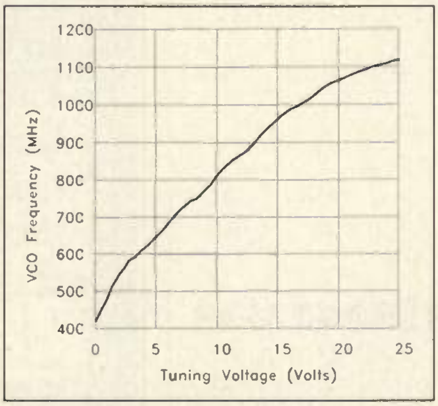

Figure 12. Tuning curve for the 500-1000 MHz VCO.

The author would like to thank Dr. Ulrich Rohde for his support and insightful help on phase noise measurements and theoretical considerations on the oscillators, and Mr. Shankar Joshi, Chief Engineer at Synergy Microwave, for his constant support and technical help on finalizing the conceptualized units.

For more information on this new VCO family, please call Synergy Microwave at (201) 881-8800, or circle Info/Card #250. **RF**

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About the Author

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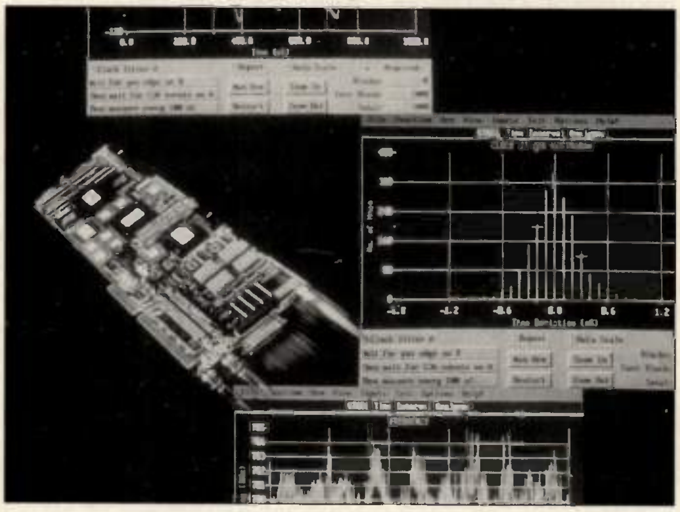
INFO CARD 48

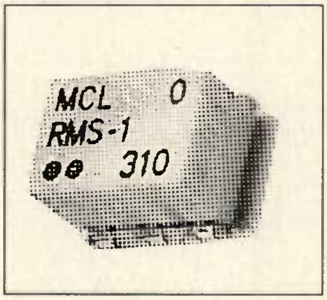
Time Interval Analyzers

Guide Technology has introduced a line of time interval analyzers and modulation domain analyzers. Named the GT650 Series, these AT/386 plug-in boards are the latest addition to Guide Technology's line of high performance time and frequency instruments. There are five models in the series — three with two channels and two with 16 channels. The two channel models have a frequency range from DC to 400 MHz (2 GHz optional), with measurement rates from 2 MS/s to 10 MS/s, and single-shot resolution from 100 ps to 2 ns. The 16 channel models have a frequency range of DC to 25 MHz, with measurement rates of 2 MS/s and 25 MS/s, and single-shot resolution of 20 ns. Being on a PC, these time interval analyzers offer significant advantages in performance and value. The software for these instruments is far more complete than what is customary on benchtop IEEE-488 instruments. Programs are provided for high speed measurement directly to PC memory, with extensive analysis capability. All models come with a DOS-based software package providing a virtual instrument front-panel on the PC's monitor for interactive benchtop use. An MS-Windows virtual front panel is expected by the fourth quarter 1993. The GT651 is priced at $13,500.
Guide Technology, Inc.
INFO/CARD #249



Surface Mounted Mixer

RMS-11 is a double balanced mixer which spans a very wide bandwidth, 5-1900 MHZ, in a compact 0.25 × 0.31 × 0.2 inch package. Newer design techniques have been adapted to widen the bandwidth to 1.9 GHz and maintain good LO-RF and LO-IF isolations. Average values of LO-RF and LO-IF at 1.9 GHz are 26 and 28 dB and at 1 GHz are 29 and 31 dB. A single mixer can satisfy the requirement of both cellular and PCN bands. Conversion loss is 7.2 dB in the 800-1000 MHz band and 8.5 dB in the 1800-1900 MHz band. The



mixer has an all ceramic body, thus matching temperature coefficient of different parts, and it uses all-welded construction. The part is also available in leaded version as LRMS-11X. Both parts can be supplied in tape and reel, use ultra-rel diodes and carry a five-year warranty.
Mini-Circuits
INFO/CARD #248

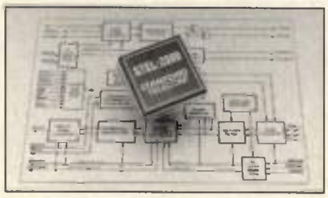
High Volume, Single Layer Capacitors

American Technical Ceramics is pleased to introduce its ATC 114 Parallel Plate Capacitors, designed for high volume applications. These high quality, very low cost, single layer capacitors

have an operating frequency range to 50 GHz. The ATC Universal Barrier Termination's superior bond strength ensures bondability and solderability, while providing versatility in attachment techniques. All ATC 114 capacitors meet all relevant and applicable industry and military specifications. Three stock sizes are offered: 25, 35 and 50 mils square, and values from 1.0 pF to 270 pF at 100 WVDC are offered. Kits containing five values, ten pieces per value are available for $29.95, and kits containing ten values, ten pieces per value cost $39.95. Orders are shipped within 48 hours.
American Technical Ceramics Corp.
INFO/CARD #247

Communications Signal Processor

The ASIC and Custom Products Division of Stanford Telecommunications announces that first-silicon prototypes of the STEL-2000 digital, fast acquisition, spread spectrum communications processor have arrived and will soon be available for Beta-site sampling. The STEL-2000 is a single CMOS ASIC device which performs all the digital processing required to implement the baseband functions of a fast acquisition direct sequence, spread spectrum, radio link. Capable of transmission in bi-phase shift keying (BPSK) or quadri-phase shift keying (QPSK) modes, it operates at 10 MChips per second in transmit and receive modes. A proprietary acquisition processing technology used in the STEL-2000 permits acquisition of



bursts of data with a single symbol preamble, making the modem extremely efficient when operating with short bursts. All parameters in the device are fully programmable, allowing it to operate over a wide range of conditions in a large number of applications.
Stanford Telecommunications, Inc.
INFO/CARD #246

Load Pull and Noise Measurement

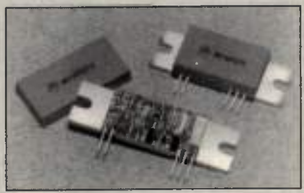
Focus Microwaves introduces the Microwave Tuner System (MTS), a fully programmable load/source pull measurement

system for noise and high power applications in the 800 to 3000 MHz frequency range. The basic configuration includes two programmable tuners (max. VSWR > 10:1, RF resetability > 50 dB) with SMA connectors (other connectors optional), an IBM-PC compatible tuner controller and GPIB interface and software for tuner, setup and test fixture calibration, noise figure and noise parameter measurement, load and source pull of power, gain, efficiency, intermod and JDC spectral performance. Load/source pull data is converted to ISO contours over load/source impedance charts. Options include: extended frequency from 750 to 4200 MHz, software for test fixture, PC compatible two channel analog interface, calibration services, and two day training. The MTS basic system costs $19,900. Delivery is two weeks ARO.
Focus Microwaves, Inc.
INFO/CARD #245

Product Spotlight: Amplifier Modules

Digital Cellular

Motorola's MHW926, MHW927A and MHW927B are UHF, linear power amplifier modules designed to serve the needs of the new United States Digital Cellular (USDC) system. The devices deliver six watts



average power, operate with a 12.5 V supply and cover the 824 to 849 MHz band. All feature 50 ohms Zin and Zout and have a third order intermodulation distortion specification of –29 dBc. Prices are $55.44, $56.25 and $56.25 for the MHW926, MHW927A and MHW927B, respectively.
Motorola Semiconductor
INFO/CARD #244

High OHIP2

The Locus model RF-2286A provides an output second harmonic intercept point, OHIP2, of +60 dBm minimum over a wide temperature range with less than 4.0 W of DC power. The unit utilizes GaAs MMIC/MIC thin film construction and measures 1.5 x 2.5 x 0.350 inches, less mounting and I/O projections.
Locus, Inc.
INFO/CARD #243

Modular Series

Each member of Comlinear's Modular Series amplifiers has been optimized in a particular performance area to provide the features best suited to specific applications. Features include true DC coupling, 160 MHz to 1.1 GHz bandwidths, fixed or flexible gain and I/O impedance, BNC or SMA female connectors, 10 dBm to 27 dBm max. power outputs, and –25 to +85 degree C operating range.
Comlinear Corp.
INFO/CARD #242

Low Noise

Model AU-1415 is a low noise amplifier operating over the frequency range of 1 to 400 MHz. Typical gain is 45 dB and typical gain flatness over the entire band is less than ±0.5 dB. Noise figure is 1.6 dB max. and 1.4 dB typ. at 400 MHz. Internal voltage regulation allows operation at any DC voltage from +15 to +30 V, with nominal current draw at 135 mA.
Miteq, Inc.
INFO/CARD #241

Cascaded Power Amplifier

The model A2P2520 thin-film cascadable amplifier operates over the frequency range of 100 to 2500 MHz with a minimum gain of 34.5 dB (0 to 50 degrees C). Noise figure is typically 3.2 dB. Typical output power at the 1 dB compression point is +27.5 dBm. typical third-order intercept point is +40 dBm. Operating current is 370.0 mA at 15 V.
Cougar Components
INFO/CARD #240

Digital Cellular Receive

The QBS-141 and -142 were specifically designed for digital cellular receive base stations. Both devices have less than 1 dB noise at +25 degrees C, two redundant RF signal paths, 90 degree hybrid combiners for soft-fail, reverse voltage protection to 100 V, and a DC power input fuse protected to 2.0 A.
Q-bit Corporation
INFO/CARD #239

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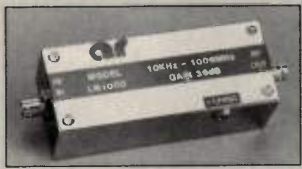
TECH * KNOWLEDGE

Broadband Pre-Amplifier

Operating within a bandwidth of 10 kHz to 1000 MHz, the AR Model LN1000 offers a combination of low noise (3.5 dB typ., 4.5 dB max., from 2 to 1000 MHz) and high gain (30 dB minimum) for applications where



system noise level must be reduced while retaining existing components. Output power with a –18 dBm input signal is +11 dBm at less than 1 dB compression; harmonic distortion is –20 dBc max. at rated output. Price is $700, delivery from stock.

Amplifier Research
INFO/CARD #238

80W Power Amplifier

This RF power amplifier has an output power level of 80 W CW from 100 to 500 MHz. The unit can be customized to deliver 100 W output over the frequency band from 225 to 400 MHz. Model 500- 100-80-35A delivers high efficiency in a small package size. The unit measures 3.2 x 2 x 1 inches (10 dB size) and 4.84 x 2 x 1 inches (for 35 dB or gain).

LCF Enterprises
INFO/CARD #237

MMIC Amplifier

Designed for operation at 2 to 18 GHz, this physically small 1.65 inch long unit provides 22 dB of gain and a 1 dB compressed power of 18 dBm. Noise figure is 7 dB max. up to 14 GHz and 8 dB maximum at 18 GHz. Maximum RF input survivability is +20 dBm. Optional input RF limiters are available to protect the amplifier against higher levels. Field removable SMA connectors are provided for drop-in applications.

AML, Inc.
INFO/CARD #236

40/45 dB DLVA

Model LVDN-0518-50 is a 0.5 to 18 GHz, truly DC-coupled, 40/45 dB detector log video amplifier (DLVA) with noise immunity circuitry. The rise time is 20 ns max. and the recovery time is 150 ns typically. The TSS is –42 dBm minimum with linearity, frequency, and baseline accuracy of ±1.0 dB. The LVDN-0518-50 operates from ±9 V to ±18 V at 75 mA. Size is 2.3 x 2.3 x 0.45 inches.

American Microwave Corp.
INFO/CARD #235

DISCRETE COMPONENTS

Variable Capacitor

A new PTFE high-voltage variable capacitor boasts 0.025 percent linear resolution. Suitable in motor driven control applications, this capacitor features roller bearing construction with an anti-backlash leadscrew. These durable capacitors contain no glass or bellows.

Polyflon Co.
INFO/CARD #234

Phase Shifters

The B4QF series of 90 degree phase shifters includes parts available from 1 to 500 MHz. Typical insertion loss is 3 dB, amplitude unbalance is less than 1 dB, impedance is 50 ohms, and phase change is typically within 1 to 3 degrees of 90 degrees. The series has a low profile height of only 3.5mm and a 6mm square footprint.

Toko America, Inc.
INFO/CARD #233

SIGNAL PROCESSING COMPONENTS

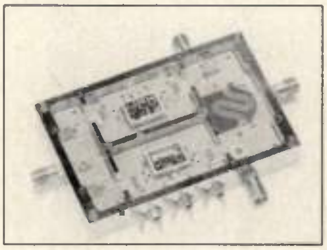
Filter Bank

Model 13SFBX-1000/3400P is a 13 channel filter bank covering input frequencies of 1.0 to 3.4 GHz. Insertion loss is 6.0 dB max., while rejection is 60 dB at ±200 MHz from center frequency. The 3 dB bandwidths are 5 to 10 percent of center frequency, and switching time is 1 us max.

Eastern Multiplexers, Inc.
INFO/CARD #232

I&Q Demodulator

Merrimac model IQM-35-1575 is a complex I&Q demodulator designed to cover the 1575 MHz



GPS band. The design includes high speed electrically switched calibration circuits, allowing precise setting of quadrature under a wide range of conditions.
Merrimac Industries, Inc.
INFO/CARD #231

Diplexer

K&L's dielectric resonator diplexer channelizes two sets of frequency bands (824 to 849 MHz and 869 to 894 MHz) to one common port. Maximum VSWR is 1.4:1, insertion loss is ≤ 3 dB across the entire band and ≤ 2 dB at mid-band. Its dimensions are 3.5 x 1.33 x 0.8 inches.
K&L Microwave, Inc.
INFO/CARD #230

Isolator

Trak Microwave introduces the first of their MICpuck™ family of isolators operating in the L-band. The magnetically shielded, metal encased isolators operate from 1.37 to 1.44 GHz, have isolation of 20 dB min., insertion loss of 0.5 dB max. and maximum VSWR of 1.2:1. Connections to the 25 mil substrate are gold and the termination is 1 W, although a 15 W termination is available for use with proper heat sinking.
Trak Microwave Corp.
INFO/CARD #229

Caseless Coupler

Sage Laboratories has developed an octave band, 10 dB, drop-in, caseless coupler to complement its family of Wirepac™ 90 degree hybrids and couplers. The model GCB6 series provides excellent coupling flatness and directivity greater than 20 dB over octave bands from 250 to 4000 MHz. These couplers are only 0.145 inches in diameter but handle 400 W average power at 300 MHz.
Sage Laboratories, Inc.
INFO/CARD #228

110 dB SPDT Switch

Model SW-2187-1DS is a reflective PIN-diode pulse modulator/switch with a frequency range of 0.5 to 18 GHz, video transients of ≤ –70 dBm, and isolation of 110 dB from 1 to 18 GHz. The switching speed is 50 ns and the insertion loss is 4.5 dB max. Control is single-ended ECL. Power supplies ±5 V at 50 mA. Size is 0.792 x 1.91 x 0.50 inches.
American Microwave Corp.
INFO/CARD #227

Isolator

An isolator from McManus Microwave measures only 1 x 1 x 0.4 inches and operates from 395 to 425 MHz. Isolation is 20 dB

min., insertion loss is 0.6 dB max., and VSWR is 1.20 max. The input connector is a 0.040 inch dia., 0.080 inch long glass feed- through pin, and output connector is an SMA female.
McManus Microwave
INFO/CARD #226

SUBSYSTEMS

DSP Development

The WBC25 DSP Workbench was designed for benchtop development, yet is compact enough for embedded applications. The DSP Workbench is based on the TMS320C25 50 MHz DSP and provides a full complement of

memory and I/O peripherals. Priced at $695, the Workbench package includes Micro Platforms' Direct Access Kernel and development software compatible with MS Windows 3.1.
Micro Platforms
INFO/CARD #225

Exciter

The model DE4030 agile exciter is a multi-mode radio exciter suitable for communications, ECM and comm simulation applications. It covers 100 kHz to 150 MHz in 2 Hz steps and can tune between any two frequencies in this range in 1 microsecond. Mechanically, the exciter is a board set consisting of two, 9 x 9 inch 6U VME cards.
Digital Radio Corp.
INFO/CARD #224

Fiber Optic Antenna Link

United Technologies Photonics has announced a pedestal-mounted multichannel fiber optic link for antenna remoting applications. It features higher optical output power, higher dynamic range and lower noise than currently available fiber optic links, plus improved bandwidth and stability. The system can transmit a multichannel signal through a single fiber.
United Technology Photonics, Inc.
INFO/CARD #223

1 kW HF Transmitter

Harris RF Communications' new lightweight, compact 1 kW HF transmitter includes a microprocessor-controlled exciter and 1 kW power amplifier and power supply. The enclosure is a standard 19-inch metal rack cabinet. The RF-1140A covers the 1.5 to 29.9999 MHz range in 10 Hz increments, with frequency agile tuning.
Harris RF Communications
INFO/CARD #222

CABLES & CONNECTORS

Low Impedance Cable

The LZ series of very low impedance microstrip transmission lines is designed for transmitting high current (to 100 A), fast rise time (to 1 ns) pulses in low impedance applications. These unique lines are available with characteristic impedance values of 1, 2, 3, 6 and 12 ohms and are a standard length of one meter. Price is $290.00 per meter, with delivery from stock.
Avtech Electrosystems, Ltd.
INFO/CARD #221

Micro-Coaxial Connectors

A 50 ohm micro-coaxial contact designed for high speed data transfer has been introduced by Harting Elektronik. The twin contact is designed to be used in the IEC 1076-4-2 2.5mm high density connector system. The contacts are rated to 2 GHz, have VSWR of 1.25:1 and crosstalk attenuation of 33 to 41 dB, depending on PCB material.
Harting Elektronik, Inc.
INFO/CARD #220

Surface Mount Connector

A new surface mount miniature connector is available from Huber+Suhner. Designated the SMD-MMCX, this connector is designed to operate either as a right angle or vertical pc board launch and has excellent electrical specifications up to 6 GHz. The connectors are available as single units or in tape and reel.
Huber+Suhner, Inc.
INFO/CARD #218

High Velocity Cables

Miniature high velocity cables from Storm Products offer low attenuation and capacitance. The cable O.D. sizes range from 0.050 to 0.155 inches with a typical velocity of 83 percent. A low density PTFE dielectric and vari-

ous shield configurations result in reduced attenuation and reduced crosstalk.
Storm Products Co.
INFO/CARD #217

Coaxial Adapters

A new series of coaxial adapters from Pasternack Enterprises includes C female to BNC female bulkhead, 75 ohm N male to 75 ohm BNC female, 75 ohm BNC male to 75 ohm BNC male and 75 ohm N female to 75 ohm BNC male types. All adapters have brass nickel plated bodies, PTFE insulation, silver plated contacts and an operating temperature of –65 to +165 degrees C.
Pasternack Enterprises
INFO/CARD #216

Semi-Rigid Cable Replacement

T-flex coaxial cables from Times Microwave Systems were developed to replace RG-402 and RG-405 semi-rigid cables in both new and existing applications. These flexible cables use the standard connectors that are currently used with RG-402 and RG-405 semi-rigid cables. Attenuation specifications for T-flex cables match MIL-C-17 requirements, and shielding effectiveness is greater than 100 dB.
Times Microwave Systems
INFO/CARD #215

TEST EQUIPMENT

Voltage/Power Meters

The Rohde & Schwarz URV 35 level meter is a light, compact, precision voltage and power meter covering DC to 26.5 GHz. The URV 35 has both digital and moving-coil displays. The URV 55 performs voltage measurements to 2 GHz and power measurements to 26.5 GHz and has an IEC/IEEE bus. Both instruments use "intelligent" measuring heads which contain calibration information for accurate measurement.
Rohde & Schwarz
INFO/CARD #214

VXI Noise Modules

Noise Com has introduced a new series of VXIbus noise generators. The VXI-7000 series covers frequency bands between 10 Hz and 40 GHz in C size modules. The unit delivers white noise with a Gaussian amplitude distribution. Noise power can be attenuated from 0 to 127 dB in 1 dB steps.
Noise Com, Inc.
INFO/CARD #213

Power Meter

The Boonton 4400 peak power meter provides 14 automatic power and time measurements of pulsed power signals from 30 MHz to 40 GHz. With a 47 dB dynamic range (57 dB in CW mode), the 4400 can capture pulses with rise times as fast as 10 ns. Waveforms can be displayed on either linear or logarithmic scales. A new "peaking-mode" uses DSP to peak-detect through 5 million continuously acquired samples. Price is $13,000.
Boonton Electronics Corp.
INFO/CARD #212

Signal Generator Option

Option 8 for the Marconi Instruments 2030/2040 series of signal generators provides RF profiling and offset capability as well as user-defined segmented sweeps. RF output profiling provides output levels as a function of frequency. Segmented sweep allows generation of up to 10 sweep bands, with each band potentially having different output levels, start and stop points and frequency resolutions. Base price of Option 8 is $1500.
Marconi Instruments
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Transistor Biasing Fundamentals

By Gary A. Breed
Editor

This tutorial is a basic one; covering the requirements for biasing bipolar junction (BJT) and field-effect (FET) transistors in various circuit configurations. Biasing for both small-signal and power transistors is covered. Note that this review is not intended to teach bias calculations or actual circuit design — it is meant to remind engineers of the basic performance factors that are controlled by transistor bias.

Biasing is a key element in circuit design, with some specific requirements in RF circuits. Many requirements are common to all AC circuits, while a few RF-specific techniques must be employed to achieve the desired performance.

The simple purpose of biasing is to establish a quiescent operating point. This is done to assure that the desired output voltage/current swing occurs with the given input voltage/current swing. The first task in determining bias circuit design is defining the input-output relationship.

Classes of Operation

Transistors have transfer functions which have four regions, which are easiest to describe in terms of DC. *Cutoff* is the first region, where the applied voltage (bias-plus-signal) is below the threshold at which the device begins to conduct. As the input increases, there is a *transition region* where the transistor begins conduction. In this region, the input-output relationship is not linear. The primary region of interest is the *linear region*, where a change in input results in a predictably larger change (gain) in the output. Finally, there is *saturation*, where the transistor is at its limit of voltage or current handling and the output no longer changes with an increase in the input.

Class A operation — This is the operating condition for linear amplification. The combination of bias and input signal are always within the linear region of the transistor. Nearly all small-signal amplifiers and high-linearity power amplifiers may operate in class A.

Class B operation — Class B is nearly always used in push-pull amplifiers.

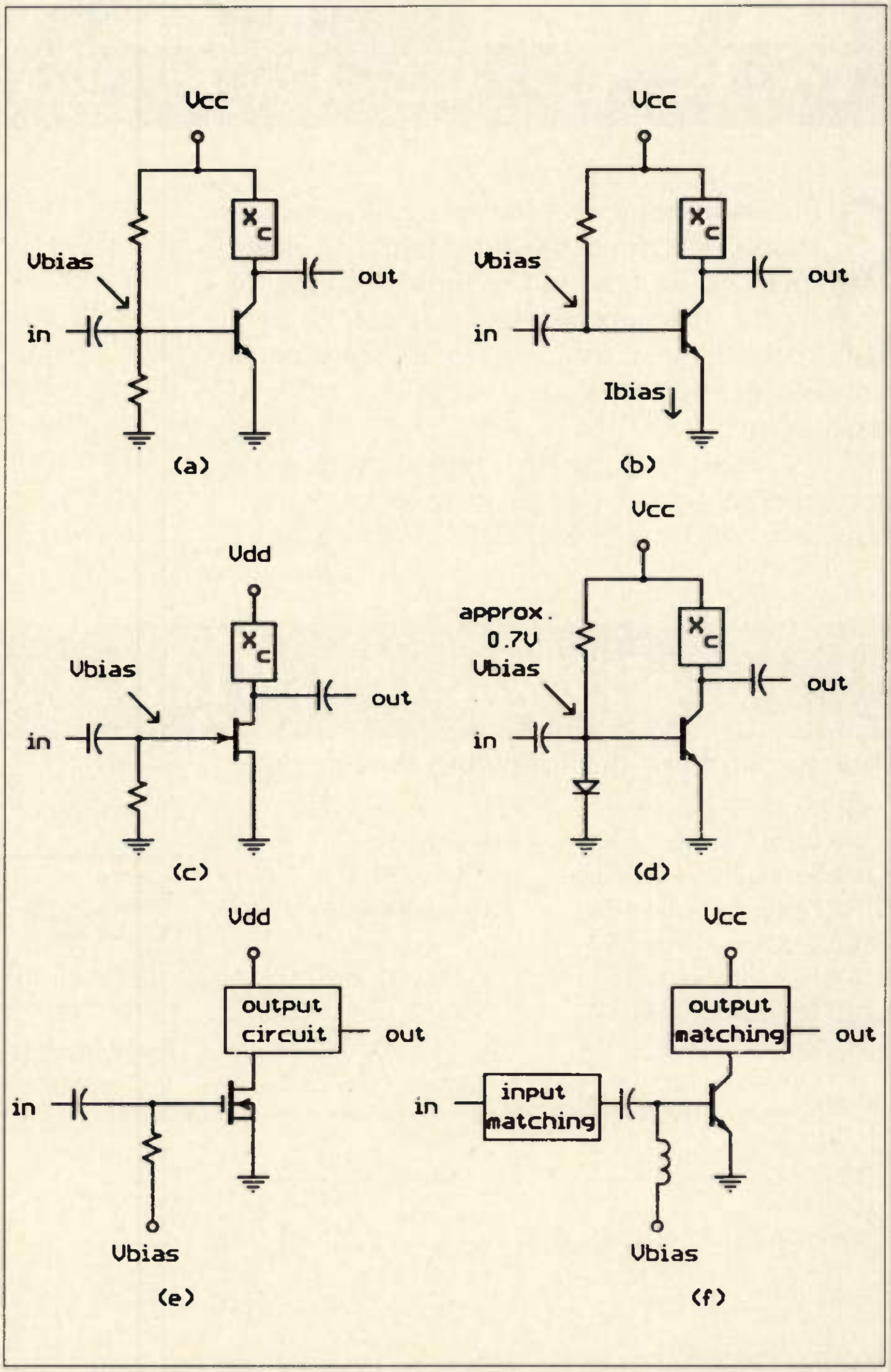

Figure 1. Several bias circuit configurations: (a) voltage divider, (b) BJT self-bias, (c) JFET self-bias, (d) diode voltage-drop bias, (e) resistor isolated bias, and (f) RFC isolated bias.

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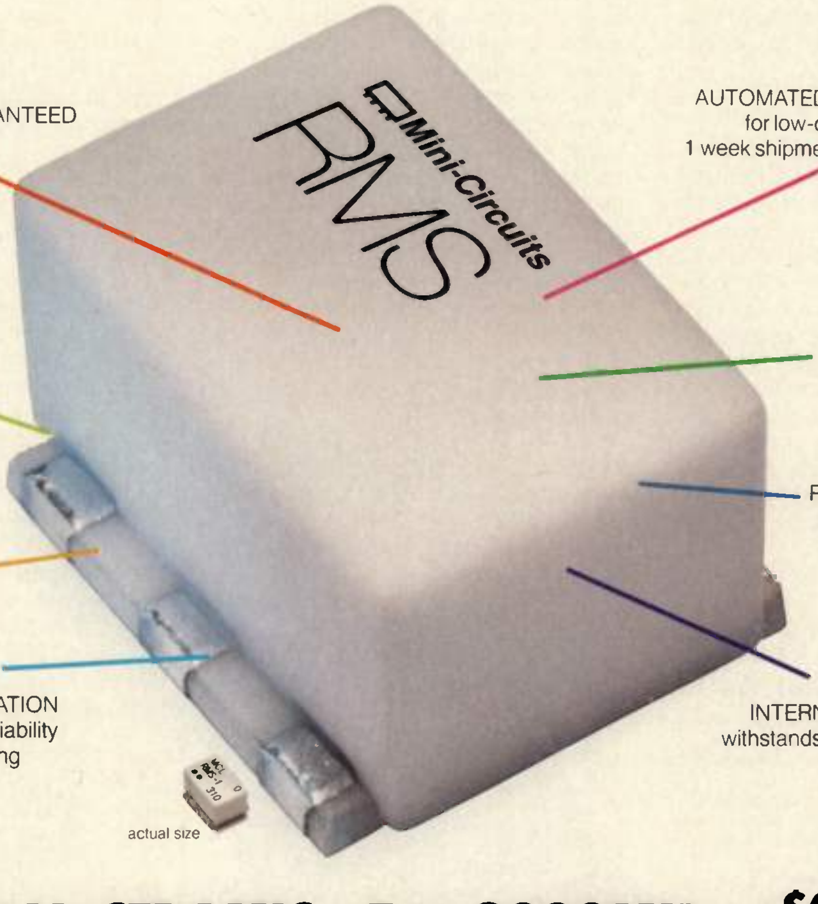

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Each device operates linearly for half a cycle while the other device is in cutoff, and the alternating contributions are combined at the output into a single linear signal. The transition region causes some problems in the changeover from one device to the other, so most push-pull amplifiers are biased a little bit toward class A, resulting in a hybrid mode called class AB. Small-signal amplifiers rarely use class B.

Class C operation — This class only exceeds the transistor turn-on threshold on peaks of the input waveform. This mode offers efficient power amplification, but is highly non-linear. Class C is only used in power amplification where linearity is not an issue (CW, FM, or with a modulated amplifier).

Bias Circuits

Developing the appropriate DC bias voltage at the base or gate of a transistor can take many forms. Figure 1 shows some of the most common configurations, including voltage dividers, self-bias circuits (using current through the base-emitter or gate-source junction to establish the bias voltage), and external regulators.

In small-signal circuits, resistive self-bias and voltage-divider bias circuits are the most common. But in critical RF applications, and at microwave frequencies where input matching may be relatively complex, the bias voltage is often generated external to the circuit and coupled via an RF choke to the circuit. Special considerations in small-signal biasing include temperature compensation, controlling bias-dependent S parameters or stability factors, and maintaining sufficient quiescent current for maximum dynamic range. Reference 1 contains a more detailed explanation of small-signal amplifier biasing considerations.

Power amplifiers have different requirements for biasing, although the basic requirement of applying a voltage to the base or gate is unchanged. In BJT power amplifiers, low impedance inputs and high RF currents require a low-impedance bias circuit that can maintain the proper voltage in the presence of currents generated by the RF input. Low bias voltages (usually under one volt) prohibit the use of a resistive voltage divider except in low power (up to 50-watts) amplifiers. The forward voltage drop across a rectifier diode is often used, since the typical 0.7-volt drop is close to the class B bias level for BJTs. For better regulation and adjustable control over bias points (to improve linearity), a transistor regulator is also often used. This approach requires selection of a regulator capable of low voltage operation.

In power FET circuits, bias current is not much of a factor, but a variable voltage is often needed. Bias is often used in over-drive or high-VSWR protection circuits, since it can bring the transistors closer to cutoff, lowering gain, reducing current drain and power output. Setting equal quiescent currents in multiple-device amplifiers is another place where adjustable bias is needed. A low impedance bias circuit is not required, but the gates are typically driven from a low-impedance RF circuit, requiring coupling capacitors to keep the bias circuit above DC ground.

Finally, temperature compensation is a key issue. In an active bias regulator circuit, a thermistor or other thermal detector must be used to automatically track bias with temperature. The complexity (and resulting precision) of the bias regulator will vary according to performance requirements and safety functions to protect the transistors from thermal damage. More information on power transistor biasing can be found in reference 2.

Summary

The message of this short review is that transistor bias requirements can be simple, or they may be quite complex. Bias circuit design is *not* a task to be handled casually. A common result of deficient bias circuits is that the transistor circuit operates adequately under ideal conditions of gain, signal level and temperature; then exhibits unexpected behavior when placed into operation in a more realistic (and variable) operating environment. ***RF***

References

1. Chris Bowick, *RF Circuit Design*, Howard W. Sams & Co.
2. Norm Dye and Helge Granberg, *Radio Frequency Transistors*, Butterworth-Heineman.

A Generic GPIB/RS-232 Controller

By Kevin Cordle

Often times it is desirable to perform repetitive tests on a design using different stimuli and under varying enviromental conditions. The IEEE-488 instrumentation bus is an ideal way to do this, using precision test equipment equipped with GPIB capability. The "GPIBCTRL.EXE" program is a general purpose GPIB controller program which allows one to rapidly write and run programs to control GPIB equipped instruments. It includes a full screen editor with cut, paste and undo capabilities and a complete and easy to learn GPIB macro language. The user interface is written using Borland's Turbo Vision and employs a graphical-like interface, including pull down menus, dialog boxes, resizable, multiple windows and full mouse support. Anyone who is familiar with the Borland Language products will feel at home with this editor.

In addition to GPIB control, "GPIBCTRL.EXE" also includes an RS-232 capability. This allows one to control, not only IEEE-488 instrumentation but also instruments controlled via serial communication. GPIB and serial control statements can be mixed in the same macro program, allowing one to talk to both GPIB and RS-232 instruments at the same time.

Online help is available at any time from within the program. This help system makes use of hypertext techniques to provide rapid access to the various help topics included.

Hardware Requirements

"GPIBCTRL.EXE" requires an IBM or compatable running DOS 3.0 or higher with 640K of memory. Best performance will be found using a 386 or 486 although it will also run on an XT type machine.

Software Requirements

"GPIBCTRL.EXE" will work with any PC based GPIB controller card which uses character based I/O and interfaces with software through a DOS device driver loaded in the Config.Sys file. The program has been tested using both Metrabyte and National controller cards.

The DOS device driver name and any initialization string required are stored in a file called "GPIBCTRL.CFG". This must reside in the same directory from which "GPIBCTRL.EXE" was loaded. The device driver name and the optional initialization string may also be entered and saved from within "GPIBCTRL.EXE" via pull down menu "GPIB". Refer to your GPIB controller card manual for the applicable initilization procedure needed. Serial communication parameters are also saved in the .CFG file and can be changed via pull down menu "COM".

Full mouse support is provided but requires a driver compatable with Microsoft 6.0 or later. This program was written using Turbo Pascal, Ver. 6.0 with Turbo Vision by Borland. It employs object oriented programming (OOP) techniques. Any changes to the source code, GPIBCTRL.PAS, will require version 6.0 with Turbo Vision and the "Science and Engineering Tools" package by Quinn-Curtis. This latter software package contains the RS-232 Pascal unit needed for serial communication.

Macro Programming

The GPIB bus is controlled from this program by entering a series of commands, one command per line, into the editor file. When the macro is complete the program in memory can then be RUN, or STEPped through one line at a time. The syntax to send a command

```
VAR
HP8590A 18                --- define Analyzer as at
                              addr 18
BEGIN
Remote HP8590A            --- Place instr on addr 18 in
                              remote
Clear HP8590A             --- Clear it
Output HP8590A MKN        --- MKN tells Analyzer to
                              enable marker
Output HP8590A CF 120MZ   --- CF 120MZ = center freq
                              to 120 Mhz
Output HP8590A MKF?       --- MKF? tells analy. to
                              send marker freq
Enter HP8590A             --- read it off the bus
END                       --- end of program
```

Figure 1. Program example.

```
VAR
HP8590A
BEGIN
Comment                   **This is the start of the Macro**
Remote hp8590a
Clear hp8590a
Outputhp8590a mkn
Output hp8590a cf 120mz
Gosub HereIAm
Pause
END                       -- End of program
HereIAm                    ------- Start of Subroutine
Loop 3                     ---- start of Loop
Output hp8590a mkf?
Enter hp8590a
Loopend                   ---end of Loop
SubEnd                    --------- End of Subroutine
```

Figure 2. Program example illustrating loops and subroutines.

out on the bus is “OUTPUT <device addr> <device command>”. The syntax for reading data from the bus is “ENTER <device addr.>”. All data read in from the bus is returned to the Editor Clipboard where it can be edited or saved to a file. In addition to the basic OUTPUT/ENTER commands, GPIBCTRL.EXE also includes a small library of macro language commands for manipulating program execution. These commands are only used within the program and are not sent out on the bus. The following macro language commands are implemented at the present time:

BELL: Rings the PC bell to alert user.

COMMENT <string>: This statement is used to comment your macro program code. It is not sent out on the bus but simply skipped over during program execution.

PAUSE: Pauses program execution, displays a user entered text string, then waits for user interaction before continuing.

WAIT <number>: Halt macro program execution <number> of milliseconds and then continue.

JMP <number>: Jumps <number> of spaces from current macro program position to new position then continues execution.

GOSUB <name>: Branches off to subroutine located at <name> position in the file, executes the subroutine, then returns to the program positon immediately following the GOSUB command. The body of the subroutine is terminated by a SUBEND statement.

INPUT : Prompts user for input, then sends that input out on the bus. This is only used in place of the <device command> position of an OUTPUT or WRITECOM statement.

LOOP <number> : Executes a body of program statements <number> of times. The loop body is terminated by a LOOPEND statement.

RUN <program> <command line parameters>: Temporarily halts macro execution, exits GPIBCTRL.EXE, runs <program>, then returns to GPIBCTRL.EXE and continues macro execution. This feature allows one to perform tasks which are not a part of GPIBCTRL.EXE.

VAR : Marks the start of a variable block where symbolic names are designated to represent the <device addr> position of GPIB statements. This block is terminated by a BEGIN statement.

BEGIN: Marks the beginning of the main program body of statements and terminates the VAR block assignments.

END : Marks the end of the main body of programming statements. All subroutines called by GOSUB are located after this END statement. Program execution is halted after END is encountered.

Further information on these commands, their syntax and use can be found in “GPIBCTRL.DOC” or can be accessed via the help menu (F1) from within the program.

Programming Example #1

The program shown in Figure 1 controls an HP8590A spectrum analyzer. It first puts the instrument in remote, then clears its state. The next command enables the marker function, then the frequency is set and read off of the bus.

NOTE that “MKN”, “CF 120MZ” and “MKF?” are commands specific to the HP8590A spectrum analyzer and not GPIBCTRL.EXE. Refer to the programming documentation for the instrument you wish to control for the required code.

Programming Example #2

The program shown in Figure 2 is the same as #1 above but highlights the use of the LOOP, GOSUB and COMMENT macro commands to read three center frequency values from the HP8590A analyzer.

Error Trapping

One of the major advantages of using GPIBCTRL.EXE over programming individual programs using BASIC, PASCAL, etc. is that it provides ready made error trapping. This feature, in conjunction with the fact that all of the PC controller initialization is taken care of, makes this program a quick way of rapidly designing GPIB controller programs. When an error is encountered GPIBCTRL.EXE will halt, display the offending line number, and give the user a chance to RETRY, CONTINUE or ABORT program execution.

Documentation/Source Code

Included in the GPIBCTRL software package are the following files:

GPIBCTRL.TPU — Pascal unit containing help constants. This file is produced by the Turbo Pascal Ver 6.0 help file compiler from GPIBCTRL.TXT

GPIBCTRL.TXT — ASCII help file. This text is used by the Turbo Pascal Ver. 6.0 help compiler to produce the online, hypertext, help file.

GPIBCTRL.HLP — Compiled, hypertext version of GPIBCTRL.TXT.

GPIBCTRL.EXE — Compiled, executable GPIB Control program.

GPIBCTRL.PAS — Source code for GPIBCTRL.EXE.

GPIBCTRL.CFG — Configuration file read by GPIBCTRL.EXE upon initial start up. This ASCII file contains the GPIB controller card initialization strings, DOS device driver name and the RS-232 com port parameters. The initial, values are for the Metrabyte (MBC) GPIB card.

GPIBCTRL.DOC — User manual in Microsoft Word for Windows, Ver. 2.0 format

GPIB.ICO — GPIBCTRL icon for users of Windows 3.X.

The author wishes to relay a special word of thanks to National Instruments for the temporary loan of their PC GPIB controller card in testing this program. **RF**

This article is available on disk from the RF Design Software Service. See page 98 for ordering information.

About the Author

Kevin Cordle is an RF Engineer with Jerrold Communications, Inc., a division of General Instrument Corp. Currently he is involved in the design of CATV set top converters for use in Western and Eastern Europe. He holds a patent for a CATV RF switch design. His hobbies include history and programming in Pascal. He can be reached at 1330 Capital Pkwy., Carrollton, TX 75006, or by phone at (214) 323-4262.

An Isolation Probe for Oscilloscopes

By Eugene E. Mayle
R.L. Drake Co.

Virtually every consumer electronics product requires a power supply. More often than not the supply is derived from the AC line. Because of advantages in efficiency and size, off-line switching supplies are frequently chosen over line transformer types. Whether designed in-house or purchased as a subsystem, the switching supply will eventually require troubleshooting. This will most likely involve the monitoring of waveforms on an oscilloscope. For the non-isolated side this can pose considerable risk of shock. Even with an isolation transformer the chassis of the scope presents a non-insulated contact to one side of a potentially lethal source. Portable LCD based scopes are not cost effective in a lab environment where bench scopes are already in place and isolated probes utilizing optocouplers can cost more than some scopes. By exploiting fundamental RF concepts, an inexpensive and high performance isolation probe can be easily built.

This paper will describe the third of three prototypes. A summary of the development of each prototype is presented below. Prototype 1 established the basic feasibility of achieving a 5 MHz response across the isolation boundary. Using an intermediate frequency (IF) of 50 MHz, filtering requirements were eased and the viability of PN junction diodes was maintained. Transformer core and winding variations were evaluated for effectiveness. Input impedance and attenuation were set at 19 kohms and 1/19 respectively. Problems to be resolved included large ripple in the scope's displayed response, spurious oscillations in the bipolar Colpitts local oscillator (LO), an erratic DC offset, trailing edge overshoot on squarewaves, and asymmetrical swing compression.

Prototype 2 included mechanical refinements such as a circuit board mounted 9 V battery and switch, a three position input switch, and a pivoting rigid ground tip. The isolation gap on the board was increased, the LO was changed to a transformer coupled JFET type, and a static DC offset adjustment was added. Input impedance was selectable between 500 kohms (DC), 20 kohms (DC), and 20 kohms (AC). Trailing edge overshoot had vanished but the reason was unclear. Ripple and offset problems persisted. Parasitic capacitance in the input switch caused response variations between selections.

In prototype 3, the input switch was eliminated by using forked input paths of 22 kohms and 440 kohms. The battery was taken off of the PC board to allow slimming of the probe. The rigid ground tip was replaced with the flexible alligator clip type. The DC offset anomalies including asymmetrical swing were solved by dynamic balancing in the modulator, capacitive coupling of the isolated grounds and partial ground shielding of the input paths. Trailing edge overshoot was minimized through input lowpass filtering.

Block Diagram

As each block is described (see Figure 1), keep in mind the design goals of miniaturization and efficiency. The function of the attenuator block is to provide one of two levels of voltage attenuation, constant from DC to several MHz. The attenuator should lightly load generators of moderate impedance (1 kohm) while presenting a predictable source impedance to the modulator. Full scale CRT readings of ±400 V_p are ultimately desired.

The input lowpass filter should minimize spurious output signals in the passband by greatly attenuating input signals at or near the LO and its harmonics. Additionally, the filter should provide attenuation of the LO leaking to the input. It may also provide transient compensation.

The modulator should convert the attenuated baseband input signal to sidebands of a much higher intermediate frequency. A maximum input voltage of ±1 V_p should be converted with little distortion. Drive requirements should be minimized and synchronizable with the demodulator. Integral drive isolation is required.

The IF coupler should couple the modulator signal to the demodulator while providing galvanic isolation of high voltages (>1000V) between the same. A high immunity to external fields is desired.

The power splitter should provide equal and in-phase drive to both the

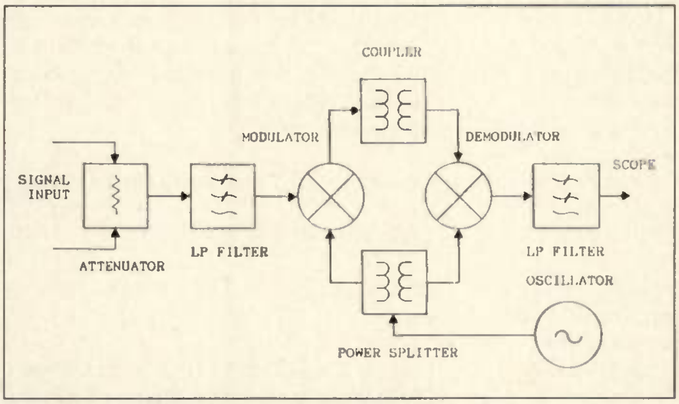


Figure 1. Isolation probe block diagram.

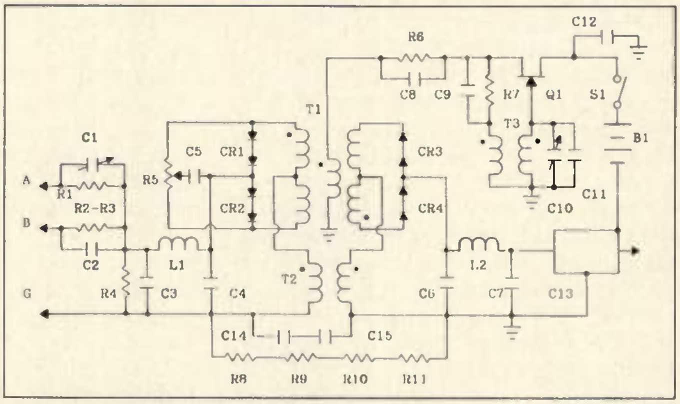


Figure 2. Isolation probe schematic.

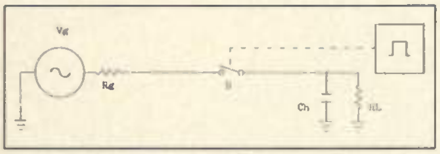


Figure 3a. Simple sample and hold circuit.

modulator and demodulator while also providing galvanic isolation of high voltages (>1000V) between the same. Loss and external radiation should be minimized. High frequency operation with nominal bandwidth and inherent impedance scaling to the LO are implied.

The demodulator should convert the double sideband intermediate signal (coupled from the modulator) to a baseband signal which proportionally duplicates the input signal. Drive requirements should be minimized and synchronizable with the modulator.

The output lowpass filter should pass signals below the cutoff frequency with minimum distortion while greatly attenuating those far above cutoff (i.e. the LO). The filter may provide frequency and transient response compensation.

The LO's function is to control the switching action of the modulator/demodulator pair. Control of the frequency, duty cycle, and efficiency is implied. The LO should operate from a 9 V transistor battery providing approximately 40 hours of operation.

Schematic

The probe's schematic is shown in Figure 2. Appendix 1 lists component specifications. Most of the components of Figure 2 can be assigned to a specific functional block in Figure 1. The remaining components are transitional — serving two functions.

Signals enter the probe through input A or B with G as the common reference. The total resistance of R2 and R3 was chosen to be 20 times greater than R1. This ratio provides good on-screen resolution of switching supply drive signals through input A and output flyback through input B. The dual inputs allow two levels of signal division without the added parasitic capacitance of a switch. Parasitic capacitance, inherent in R2 and R3 of input B, produce a zero in the probe's response which must be compensated for. This compensation is partially provided by C3 which is also an element of the input lowpass filter. To maintain equivalent frequency response through input A, variable capacitor C1 is provided.

Because the response of the lowpass

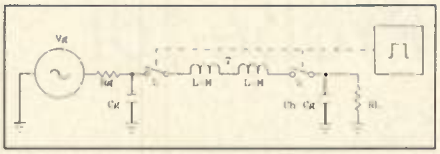


Figure 3b. S/H with floating transformer.

filter relies upon the impedance and transmission characteristics of the modulator/demodulator and coupling transformer, these circuits will be presented first.

The modulator and demodulator are identical and reciprocal circuits. Dual diodes CR1 and CR2 are used as current controlled switching elements in the modulator, while CR3 and CR4 are used in the demodulator. Separated differential secondary windings of transformer T1 provide in-phase switching of the diodes. The drive from the LO is stepped up 2:1 via the primary winding of T1. The balanced taps of the secondary windings feed 1:1 primary and secondary windings of coupling transformer T2. The returns of these windings are made to their respective grounds. The phasing of T2 is such to produce a positive output for a positive input. Toroidal cores are used for all transformers and inductors in the probe because of their shielding properties.

No matter how meticulous one is in the selection of components and the construction of this circuit, some imbalance in the modulator/demodulator will exist. Potentiometer R5 is paralleled across series diodes CR1 and CR2 with its wiper capacitively coupled to the secondary tap. This provides a means to adjust the balance of the modulator/demodulator and eliminate the subsequent DC offset.

Operation of the modulator/demodulator could be explained in the familiar terms of frequency translation, IF coupling, and synchronous detection. However, illustration of a transformation of a sample and hold circuit to the modulator/demodulator of Figure 2 will be used instead.

Figure 3a is the basic sample and hold circuit with sampling switch, S, and hold capacitor C_h. If the time constant $R_g \times C_h$ is small and the sampling frequency high, the charge on C_h proportionally tracks the level of signal generator V_g.

Figure 3b splits, S, and C_h into two parts. A floating transformer, T, is introduced between the two switches. The mutual inductance, M, cancels the winding inductance, L, and therefore this transformer ideally presents no imped-

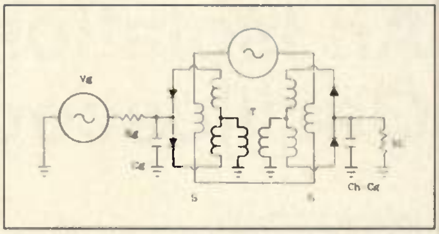


Figure 3c. S/H with isolation transformer and electronic switch.

ance to signal flow. This type of device is used in "all-pass" delay equalizers for filters. The splitting of C_h allows the transfer of charge to be more efficient by bypassing the input attenuator as a load for the modulator's sampled signal.

By grounding T in Figure 3c, isolation is created. The arrangement of the dual switches prevents a DC short for both source and load. Having been replaced by their diode equivalents, the switches operate at a near 50 percent duty cycle as the LO alternately forward and reverse biases the diodes via the drive transformer.

Because the switching spreads the signal's energy over many harmonics, the coupling transformer's bandwidth becomes crucial to coupling efficiency. Therefore, synchronous switching is required for proper operation.

The modulator/demodulator could be thought of as a transmitter and receiver pair with the coupling transformer replacing the antennas and free-space. By combining the dominant elements of diode and transformer models, a baseband model representing the modulator/demodulator and coupling transformer losses was developed. This model is depicted in Figure 4. An important test of the model's validity is that it generates the same conversion loss as the actual circuit. Therefore, conversion loss measurements became the means to calculate the model resistor values. The inductances in the composite comprise the T equivalent of the coupling transformer. The equations for calculating the T equivalent of a 1:1 transformer are given below:

L(series) = L(winding) – M(mutual ind.) (1)

L(shunt) = M(mutual ind.) (2)

M(mutual ind.) = L(winding) × K(coupling) (3)

All of these elements would change if the components were altered. Even the small amount of bias current flowing

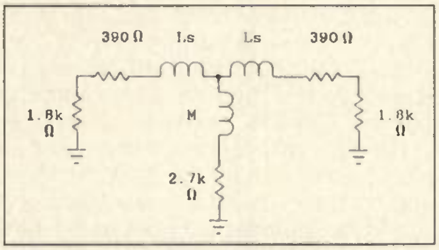


Figure 4. Modulator/demodulator baseband model.

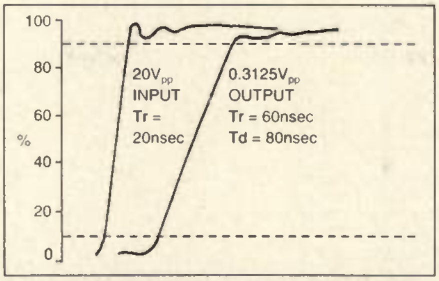


Figure 5a. Input A step response.

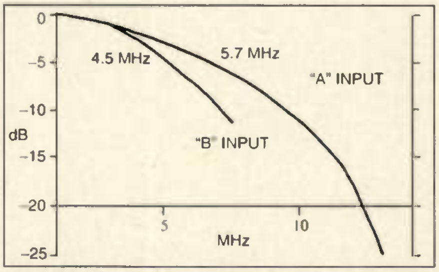


Figure 5b. Isolation probe's transmission response, inputs A and B.

through the drive winding affects the loss.

The values in the lowpass sections C3-L1-C4 and C6-L2-C7-C13 would cause severe ringing if the model was a short. The actual step response through input A of the probe is illustrated in Figure 5a. The frequency response for both inputs is plotted in Figure 5b. With only a 20 V_{pp} signal available, measurements through input B are difficult. A breadboard of the model and lowpass sections (including input attenuator) shows good correlation with the input step response.

The model is provided for those who wish to optimize the response using computer software. It should be noted that the circuit elements of Figure 2 were obtained empirically before the model was developed. Optimization must weigh the importance of the following requirements:

1. Attenuation of input signals at frequencies close to the LO.
2. Attenuation of LO leakage to the input.
3. Attenuation of LO leakage to the output.
4. Light loading of the circuit being monitored.

The force that drives this near miracle of DC transmission through an insulated boundary is the LO. The choice of oscillator type usually follows from experience after analyzing the requirements. With two series diodes in each arm of the modulator it is referred to as a class 2 mixer. Theory and experience tell us that class 2 mixers require +13 dBm of LO drive. Additionally, minimization of spurious outputs from mixers require low LO harmonic distortion.

The transformer coupled JFET design of Figure 2 was found to satisfy these goals. With transformer feedback (T3) between gate and source, a winding ratio of 3 to 2 respectively resulted in less than 5 percent harmonic distortion at +14 dBm output into 200 ohms. The oscillator draws 10.8 mA at 9 V, resulting in a DC to AC efficiency of about 27 percent. At 6 V the efficiency is about 34 percent with an output of +13 dBm. The theoretical efficiency of an oscillator meeting these goals is 50 percent. C8, C9, and C12 serve bypass functions. R6 diverts some of the bias current into the modulator/demodulator transformer T1 to affect conversion response. Oscillation frequency is determined by the primary winding inductance of T3 resonating with C10 and C11. The salient features of the LO design are low component count, frequency adjustment by changing C10, power adjustment by changing R7, small power deviation vs. supply slump, and low power drain.

Series capacitors C14 and C15 provide DC stabilization for the converted signal through the connection of the isolated grounds. Without this capacitance stray fields capacitively coupled to the balanced input are equally likely to induce current flow in the signal path as the ground path. Current flow through the signal path will be manifest as level fluctuations. Capacitance between the isolated windings of T1 and T2 is typically 1.5 pF. Adding 16 pF across the isolated grounds reduces DC fluctuations to a negligible level. Though this configuration negates true differential measurements by the probe, it greatly reduces the amount of shielding required.

Finally, resistors R8 through R11 are required to dissipate any charge that might accumulate across capacitors C14 and C15 during usage. Because the voltage rating of 0.25 W carbon film resistors is specified at 250 to 500 V, four were connected in series. The magnitude of the chain provides a short time constant (1.5 ms) and negligible current flow (11 uA) with a 1000 V differential.

Operation and Performance

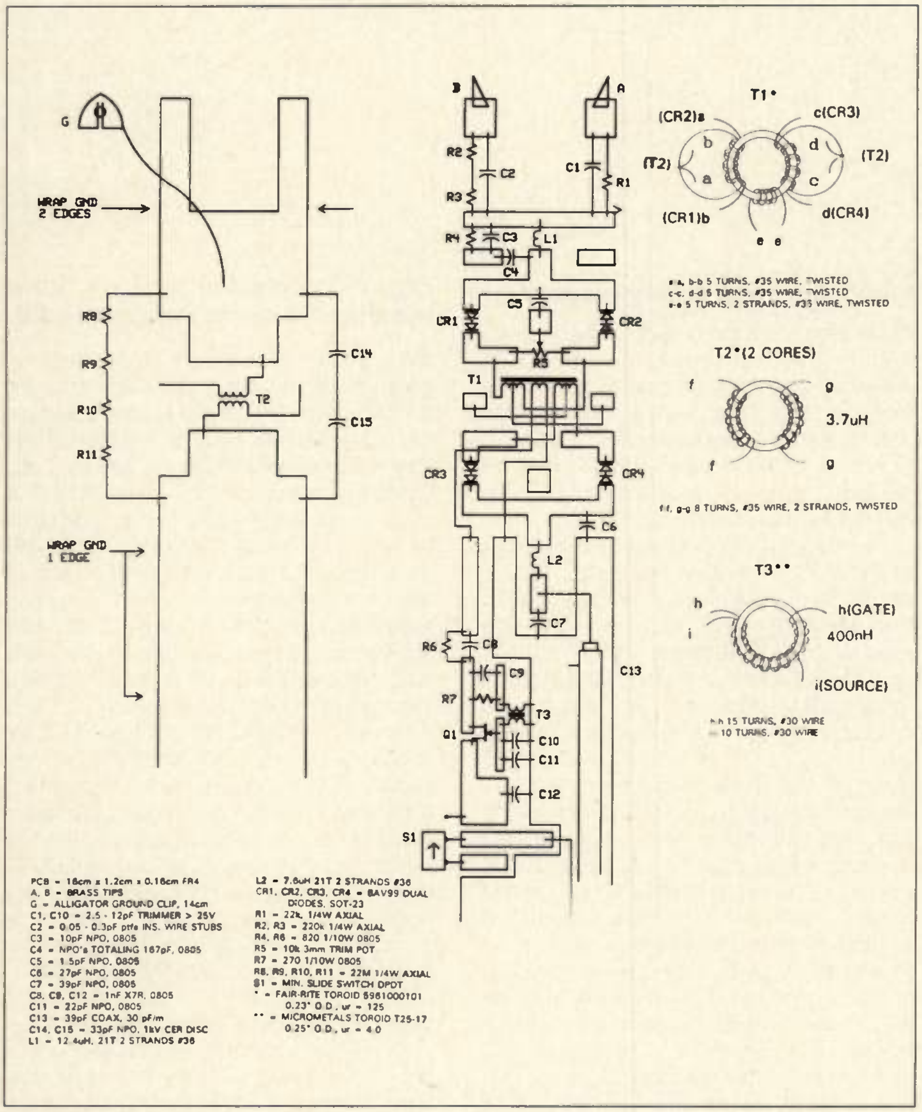
Extreme caution is advised when working with high voltages. If at all possible, the circuit under test should be de-energized and allowed time to discharge before connecting the probe. The probe should be insulated to prevent inadvertent conduction to the user or shorting within the unit under test. Adding a signal path clip-lead is advisable.

Operation is essentially the same as a standard probe. Any scope with 1 Mohm input impedance shunted by 15 to 20 pF will work. The vertical gain should be set to 20 mV/div and band limiting to 20 MHz is preferable. Connect the probe via a BNC connector to the scope. Clip on a 9V transistor battery to the supply probe using coax shield as a return and flip on the switch. After the scope has warmed up, the probe's DC offset can be adjusted. Short the probe's signal input to its ground reference; then, adjust R5 so that there is no offset between the scope's DC and GND input selections.

The transient response of the probe is

Dimensions		16.5cm × 1.2cm
Power consumption		11mA @ 9V
Sampling frequency (adjustable)		48MHz ± 3MHz
Load requirements		1MΩ // 15-20pF
	INPUT "A"	INPUT "B"
Impedance	22kΩ // <5pF	440kΩ // <0.3pF
–3dB bandwidth	5.7MHz	4.6MHz
Overshoot	<1%	<3%
Nonlinearity		
<1%	±20V_{pp}	±400V
<5%	±30V_{pp}	--------
Conversion gain	15.6mV/V	391uV/V
LO leakage	<–60dBm	<–75dBm

Table 1. Isolation probe specifications.



Appendix 1.

best set by using a 10 to 20 V_{pp} squarewave at 400 to 500 kHz. Apply the signal to input A of the probe and adjust C1 to obtain the best compromise between minimum overshoot and rise and fall times. Then apply the signal to input B and do the same thing using C2. Unless a larger signal is available the vertical scale may need a temporary adjustment. C2 in the author's probe was realized by two short lengths of PTFE insulated bus wire soldered so as to overlap. They were stiff enough to retain adjustment made by prying with a plastic flat blade. If a signal generator is available you may want to check the frequency response. For the majority of uses the transient response should be the major concern.

Though not necessary, it is recommended that the scope's display be calibrated to obtain a convenient magnitude of volts per division. I suggest 2.5 V/div through input A which results in 50 V/div through input B. This is done by applying a reference 10 V signal (DC or AC) across input A and the ground and then adjusting the vernier control on the scope's vertical gain selector. The scope's vertical position control may need adjustment after this step. Table 1 lists the specifications of the probe as obtained in the author's prototype.

Monitoring 500 kHz drive and flyback waveforms using the probe results in near identical waveforms as when using an isolation transformer with standard probes. The reference lead of the isolation probe should always be connected to the "ground" of the circuit under test when accurate high frequency waveforms are to be monitored. For near sinusoidal waveforms this is not required; though the circuit will be loaded by the 16 pF/88 Mohm coupling network C14-C15 and R8-R11.

The applications for isolated monitoring of signals are endless. With proper scaling of the input attenuator this probe would be suitable for both audio and video equipment work. A flourish to the design can be had by adding an AlGaAs LED (with series resistor) in shunt with a recalculated value of source resistor in the LO. This would serve as a power-on indicator. These LEDs require only a milliamp of current, would pose no reduction in efficiency, and might prevent unnecessary battery drain. It is advisable to maintain a visible gap between the isolated windings of T1 and T2. Coated magnet wire is used for all transformer and inductor windings called out in Appendix 1.

Conclusion

An economical high performance isolation probe has been presented. Though conceived for switching power supplies the theory and background development will allow the reader to modify the design to suit his or her own applications. It is believed that optimization could extend the bandwidth 25 to 50 percent while maintaining the same overshoot. By increasing the sampling frequency even greater performance is possible.

While hybrid isolation amplifiers are commercially available, bandwidths are typically 100 kHz or less. We all benefit when technology is widely and properly used both in economy and quality; but, this is possible only when knowledge is shared. ***RF***

References

1. W.H. Hayt, Jr., Jack E. Kemmerly, *Engineering Circuit Analysis*, McGraw Hill Book Co., New York, 3rd ed., 1978, pp. 498- 503.

About the Author

Eugene Mayle is a senior design engineer at the R.L. Drake Company. He received his BSEE from General Motors Institute in 1981 and his MSEE from the University of Dayton in 1986. He can be reached at (513) 866-2421.

Broadcast Radio Uses Home-Made Components

By Jim Escoffier
NASA

This 1992 contest entry describes an AM radio receiver that combines a detector using early-1900s technology with an IC audio amplifier of current design.

The circuit described here (Figure 1) uses a coherer for signal detection. It is probably more appropriate for a crystal receiver than a coherer, but this was done to make it recognizable by today's engineers. In a coherer circuit, a tuned front end was not generally used. The closest thing to front-end selectivity would be a resonant antenna, and that would probably only happen by accident. Also, audio amplification was hardly necessary as the sound was usually produced by a doorbell. The integrated circuit amplifier was used simply to include the oldest and newest of technologies.

Real coherer circuits tended to be somewhat less sophisticated, consisting of only a battery, the coherer and a doorbell in series, responding only to on-off keying (Figure 2). By using mechanical feedback for decohering, some resemblance to the original modulation might be maintained but the resulting audio would probably not be high fidelity.

An antenna and earth ground were connected to the two terminals of the coherer. A sufficiently strong signal would cause arcing through the powdered metals of the coherer, causing the particles to stick together and complete the circuit through the bell. Once stuck, or cohering, current flowed and the bell rang until something interrupted the circuit, or decohered it. Decohering was done with a second bell mechanism called a tapper, usually with a second battery, which continuously tapped the body of the coherer sufficiently violently to cause the particles to separate. Later implementations used a single battery for both mechanisms and a sensitive relay to switch the bell current. Neither advance caused much concern among the folks developing vacuum tube detectors.

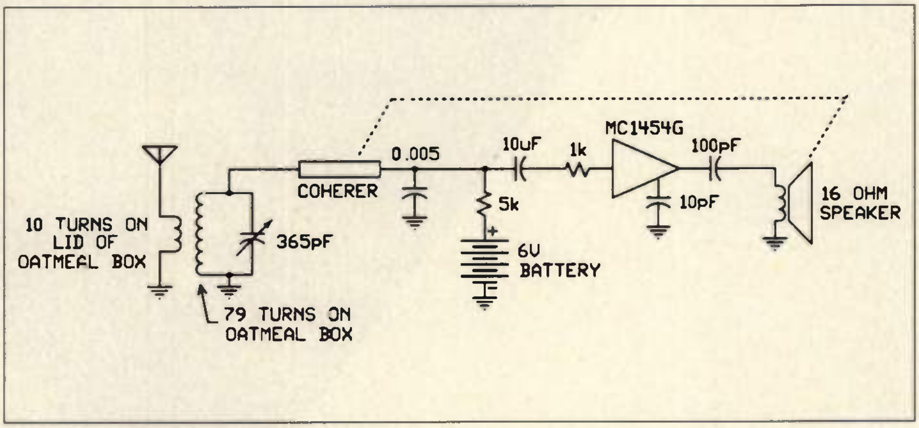


Figure 1. Circuit schematic.

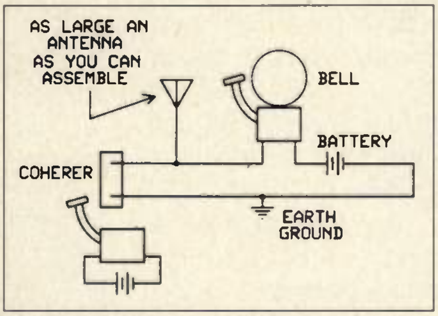


Figure 2. Typical coherer circuit.

Radio Construction

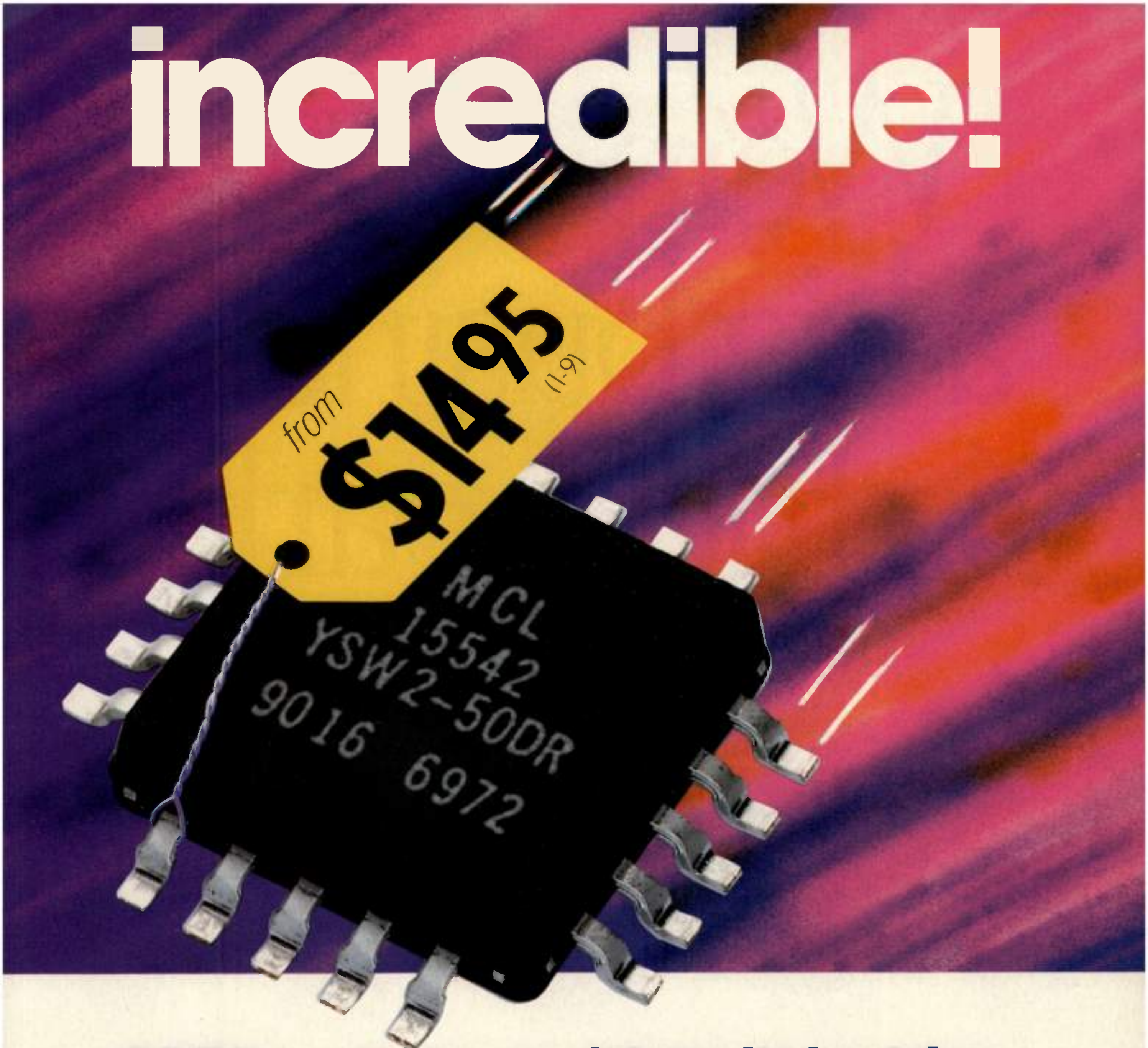
First, empty an oatmeal box. The preferred method is to dump the contents in an ordinary trash receptacle; however, if your mother won't let you do that, the contents may be eaten. Remove the top from the lid so that only a hoop remains. It should be capable of sliding down over the box. To make the coupling link, wind ten turns of no. 22 wire over the hoop and fasten the ends with glue. Wind 79 turns of no. 22 wire over the oatmeal box starting near the bottom to allow maximum adjustment of the coupling link. File down one pre-1950 silver dime and two half-inch steel washers and place the filings in a small plastic vial to make the coherer. To form the contacts, drive two small nails through opposite ends of the vial. Decohering is accomplished by taping or gluing the coherer assembly to the loudspeaker cone, where the audio vibration separates the particles.

If desired, the battery can be removed from the circuit and the coherer replaced with a crystal. The crystal can be made by sprinkling a little powdered sulphur over a spot of melted solder. A few seconds of feeling around on the residue of the sulphur with a catwhisker will produce a couple of sensitive spots. Make the catwhisker with almost any fine stiff wire. You can unwind a spare wirewound resistor to get wire for catwhiskers. As long as the resistance is fairly low it works just fine.

Bolt everything down on a scrap of wood with wood screws, and wire everything together with long wires with little curlicues on the ends, just like the old days. Extra credit will be given for using double cotton covered solid wire. Using a real diode or a PC board will be considered cheating (except for the amplifier stage — SMT may be used for that). ***RF***

About the Author

Jim Escoffier is presently working for NASA at the Kennedy Space Center, doing electronic design for Space Shuttle prelaunch checkout. He received a BSEE from Louisiana State University in 1967, but learned about radio from his father, who really did build and use coherer circuits in the 1910s. Jim's address is 1615 Saturn St., Merritt Is., Fla. 32953.



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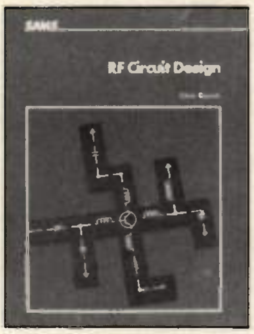

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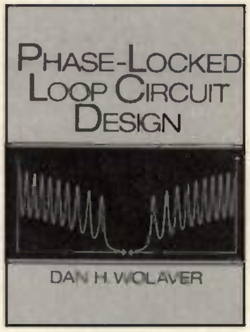

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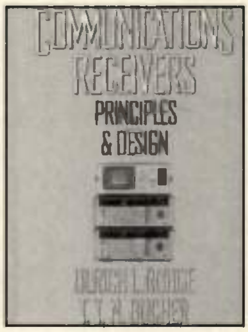

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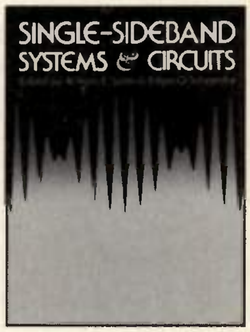

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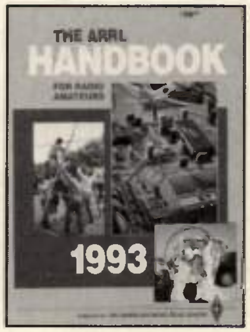

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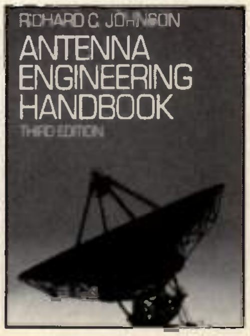

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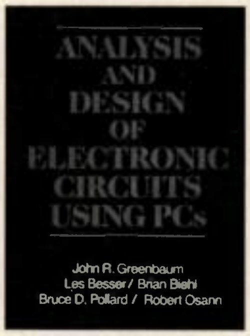

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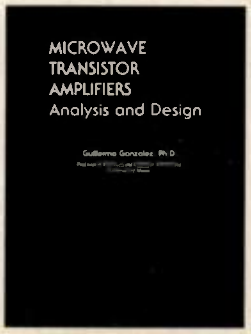

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RF books

Radio Frequency Transistors
By Norm Dye and Helge Granberg
Published by Butterworth-Heinemann, 1993

The preface says it succinctly, "This book is about radio frequency (RF) transistors." Any engineer who works with RF transistors should have this book. This is not a theoretical treatise, not a teaching text — it is a reference book and an applications guide. The emphasis is on power amplifiers, but small-signal design is also well-covered.

Radio Frequency Transistors begins with five chapters devoted to background information. "Understanding RF Data Sheet Parameters" is the title of chapter 1, followed by chapters covering the characteristics of RF transistors, comparison of FETs and BJTs, basic circuit configuration, class of operation, modulation, biasing and reliability.

After a chapter on construction methods for RF amplifiers, the book gives considerable attention to power amplifier design. Chapters 7 through 12 cover nearly all possible configurations, matching and combining, computer-aided design, filtering, VSWR protection, feedback and frequency compensation. Authors Dye and Granberg show their considerable experience in power amplifier design by presenting straightforward design information in a manner that is clear, complete, and not the slightest bit intimidating.

Although only one chapter is devoted to small-signal design, that chapter is the longest in the book. There is plenty of data on gain, matching, stability and noise, all illustrated with examples.

Phase-Locked Loops: Theory, Design, and Applications (Second Edition)
By Roland E. Best
Published by McGraw-Hill, Inc., 1993

The second edition of this book includes major updates and new material not included in the 1984 first edition. Also included with this text is a disk containing a PLL simulation program. For this edition, the author has placed equal emphasis on four types of PLLs: The Linear PLL, the Classical Digital PLL, the All-Digital PLL, and the Software PLL. With rapid growth in high-speed digital circuits and extensive application of DSP, the latter two types should be of special interest to engineers involved in current RF/digital development.

In addition to the simulation program, the book contains other useful reference data. Of special note is a list of all known PLL-related components, as of late 1992. Three appendices provide references for the PLL pull-in process, the Laplace transform, and basic digital filtering.

This is a theoretical book, providing information intended to increase the reader's understanding of PLL principles. There are no circuits to be blindly (or intelligently) copied. It is, however, the most current textbook on PLLs, and is worthy of consideration for an RF engineer's bookshelf.

Array Signal Processing
By Don H. Johnson and
Dan E. Dudgeon
Published by PTR Prentice-Hall, 1993

Signals are often detected in two- or three-dimensional arrays, and must be processed using methods that account for spatial variation. This book addresses the detection of waves in space, and is applicable to optics, acoustics, and radio-frequency signals. RF applications include interferometry, phased-array antennas, sensor arrays and time-space varying apertures. This book covers signal processing techniques appropriate for such applications: beam forming, detection theory, estimation theory, adaptive processing and tracking. Appendices add reference material on probability, matrix theory and optimization theory.

This is a technical book which assumes that the reader is familiar with linear signal processing and fundamentals of digital signal processing.

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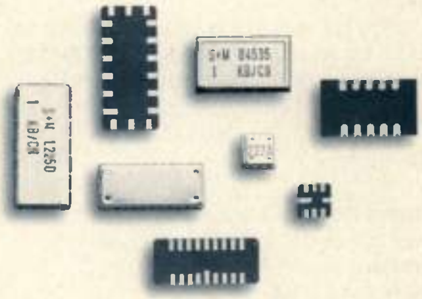

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SCI-1016

Video Op Amps: A Pretty Picture

By Andy Kellett
Technical Editor

There is more and more demand for information in the form of images. At the same time, people want better broadcast video. "Consumers are becoming accustomed to high quality video by working with computer monitors so much," says Peter Himes, Segment Marketing Manager with National Semiconductor. The demand for more and better video is pushing development of new op amps and is fueling the growth of the high-speed/video op amp market.

Specifications

The specifications which define high-speed op amps as such are bandwidth and slew rate. At the high end of the speed range is Harris Semiconductor's HFA 1100 series. Built using their bonded-wafer-isolation, silicon bipolar process, the Harris series has a unity gain bandwidth of 850 MHz.

Gain flatness is just as important as bandwidth for video applications. Most video signals carry information over a bandwidth of a few tens of MHz at the most, but gain over that bandwidth must be flat to within 0.1 dB. Using a device with a 3 dB-bandwidth several times the required 0.1 dB-bandwidth ensures adequate gain flatness.

Besides gain flatness, differential gain and differential phase differentiate video op amps from other high-speed op amps. "System specifications for differential gain and phase are typically 0.1 percent and 0.1 degree, but because the input signal loops through several gain stages, the typical requested specification for individual devices is 0.01 percent and 0.01 degrees," says Karen Cunningham, Comlinear Marketing Manger for the Signal Conditioning Product Line.

Applications

"Multimedia is kind of a nebulous term," according to Steve Pratt, Business Manager at Maxim Integrated Products, "but it is basically data, voice and imaging working around a PC." All the analog inputs and outputs to such a system must be amplified or buffered before they are either digitized or sent to an output device such as a video monitor. Many manufacturers point to multimedia as a rapidly growing market, particularly for those applications which produce professional video effects on a PC.

Although High Definition Television (HDTV) is not yet fully defined, manufacturers are able to provide some HDTV compatible devices. Manufacturers anticipate HDTV information will occupy about 30 MHz with response flat to 0.1 dB. For reasons noted earlier, the bandwidth for HDTV-compatible op amps will be about 300 to 400 MHz. This bandwidth is readily available with current technology.

Other applications are also growing. Devices using charge-coupled devices (CCDs) employ high speed op amps for signal conditioning, finding use in scanners, copiers and electronic photography. Ultrasound imagers use op amps with 20 to 50 MHz bandwidths, as do black and white surveillance video systems. As op amp bandwidths have increased, they have found use in RF applications such as driving A/D converters for digitized IF stages.

Cable driving is still a big use for video op amps. "Fully forty percent of the discrete op amps we sell are used to drive cables, and they are becoming high-powered," says Elantec Applications Manager Jay Friedman, "We have introduced a video distribution amplifier which will drive six 150 ohm loads."

Op Amp Advances

"The semiconductor industry can provide more than enough performance for 80 to 90 percent of all video applications. So it's not a case of honing and refining performance any more, it's cost reduction and moving to a higher level of integration," says Brian Mathews, Strategic Planning Manager at Harris Semiconductor.

Although the demand for reduced power consumption and reduced operating voltages will not be as urgent for video products as it will be for portable computers and radios, there is still interest in reducing those parameters. "There are strong trends for lower power just to save power, not because these applications will be portable," says Jay Cormier, New Product Marketing Manger for Analog Devices, "but some customers are driving toward a single supply because their application will be battery powered."

As for for how these changes will be made, most manufacturers agree, improved semiconductor processes will be necessary. "We're still releasing products that are state-of-the-art in some fashion, using a process that was new five or six years ago, but newer processes are really the enabling technology," says Analog Devices' Himes. However, there will be opportunity for creative design to play a role in new devices. "Chip designers here say all the advances we have made in terms of differential phase and gain, with regard to lower power consumption, is all in design," says Elantec's Friedman.

More Performance/Less Price

The high-speed/video op amp market has expanded dramatically over the last three to four years. "We're selling more every year because we are just getting into the video op amp market," says Don McGraw, Marketing Manager for Burr Brown's Analog Division, "we are building our base of products and getting them designed in." As more manufacturers have entered the market and offered more performance, prices have dropped. "Five years ago you would pay four to five dollars for a 50 MHz op amp, today you pay $1.75 in 100's for a 100 MHz op amp," says William H. Gross, Design Manger at Linear Technology Corp. Comlinear's Cunningham predicts that while prices may not drop as fast as they have the last five years, each device will contain much more performance and functionality.

Our idea of what is traditional video will certainly change in the next ten years. Editing video images on a PC or processing HDTV video will seem just as traditional as amplifying an NTSC signal today. However, high speed and video op amps will be just as important in those new applications. ***RF***

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RF literature

Pulse Amplifier Note

Microwave Technology (MwT) has generated a seven-page application note describing the performance of a 50 W pulsed power Compact Amplifier Module (CAM) operating from 960 to 1215 MHz. The module uses MwT's Solid State Triode™ power JFETs and Aluminum Nitride hybrid manufacturing capability.
Microwave Technology
INFO/CARD #209

RF Selector Guide

Motorola has released its annual RF Selector Guide and Cross Reference for 1993 (SG46/D Rev. 10). This particular guide contains the most extensive changes in Motorola's standard product offerings since 1988. To obtain a free copy of the guide, call Motorola Literature Distribution at: 1-800-441-2447.
Motorola Semiconductor
INFO/CARD #208

Revised Spec Sheets

Picosecond Pulse labs has rewritten all of its 5000 series coaxial components specification sheets to include frequency domain insertion loss and return loss plots, along with time domain transmission step response and TDR plots. An application note and spec sheet describe Picosecond's coaxial bias tees in both time and frequency domains.
Picosecond Pulse Labs, Inc.
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Spectrum Analyzer Brochure

An 18-page color brochure on Anritsu's MS2602A spectrum analyzer is available through Anritsu Sales Co. The specification, features and applications of the 100 Hz to 8.5 GHz spectrum analyzer are described.
Anritsu Wiltron Sales Co.
INFO/CARD #206

Low Noise Amplifier Data

Microwave Solutions introduces a technical data sheet of low frequency, low noise amplifiers which operate over the frequency range of 0.750 to 2.30 GHz. These amplifiers have output power of +20 dBm and noise figures from 1.8 to 2.2 dB.
Microwave Solutions, Inc.
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Design Software Report

Exemplar is offering a technical report describing software techniques for analyzing RF/microwave component interaction and for synthesizing systems. The six-page report, "PC Software Helps RF/Microwave System and Component Engineers Improve Designs" is free.
Exemplar
INFO/CARD #204

Short Form Catalog

A 124-page short form catalog, No. G-01-B, is now available from Murata Erie North America. The catalog contains detailed specifications on the company's lines of ceramic capacitors, trimming potentiometers and capacitors, EMI/RFI filters, microwave components, crystal oscillators, and more. The catalog is available free of charge.
Murata Erie North America
INFO/CARD #203

Ferrite Catalog

The 12th edition of Fair-Rite's soft ferrite catalog has been released. Four new materials are listed in the catalog. Mechanical drawings of all parts are included, and detailed electrical specifications are given for both materials and individual parts.
Fair-Rite Products Corp.
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Advertising Index

RF software

Analog Design

EEsof has announced the second release of EEsof Series IV software. Highlights of the new release include design and analysis capability for CDMA, integration with communications DSP, fiber optics, RF ASICs, and surface mount technology devices.
EEsof, Inc.
INFO/CARD #200

EM Modeling on PCs

Sonnet Software and EEsof, exclusive distributor for Sonnet products, have announced that Sonnet's electromagnetic analysis tools are available on the PC for the first time. The tools include xgeom™, em™, and emvu™ and will run on 386- and 486-based PCs using the DESQview/X™ windowing environment.
Sonnet Software, Inc., EEsof, Inc.
INFO/CARD #199

Analog/Mixed Mode Simulation

The latest version of the Intusoft ICAP/4M package for the Macintosh includes Monte Carlo statistical yield analysis, parameter sweeping and circuit optimization. These capabilities are integrated in the PreSpice module for ICAP/4M, which also includes model libraries, modeling utilities, and a netlist editor. ICAP/4M is $1575. PreSpice 3.2M is available separately for $275.
Intusoft
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I/Q Signal Simulation

IQSM software from Rohde & Schwarz allows the mathematical simulation of all digital modulation presently known. I/Q signals are computed according to user-specified modulation type, coding, baseband filtering, and bit sequence. Simulated signals are transferred to an arbitrary function generator whose output, in turn, feeds a signal generator with built-in I/Q modulation.
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Terrain Analysis Support

SoftWright's TeleTAP on-line bulletin board service provides answers to questions about their software, the Terrain Analysis Package (TAP™). Callers can instantly receive answers to commonly asked questions, leave questions to be answered and download demonstration programs. There is no charge for use of TeleTAP. Access is at (303) 344-5378 using the protocol 8,N,1 at 9600 baud or slower.
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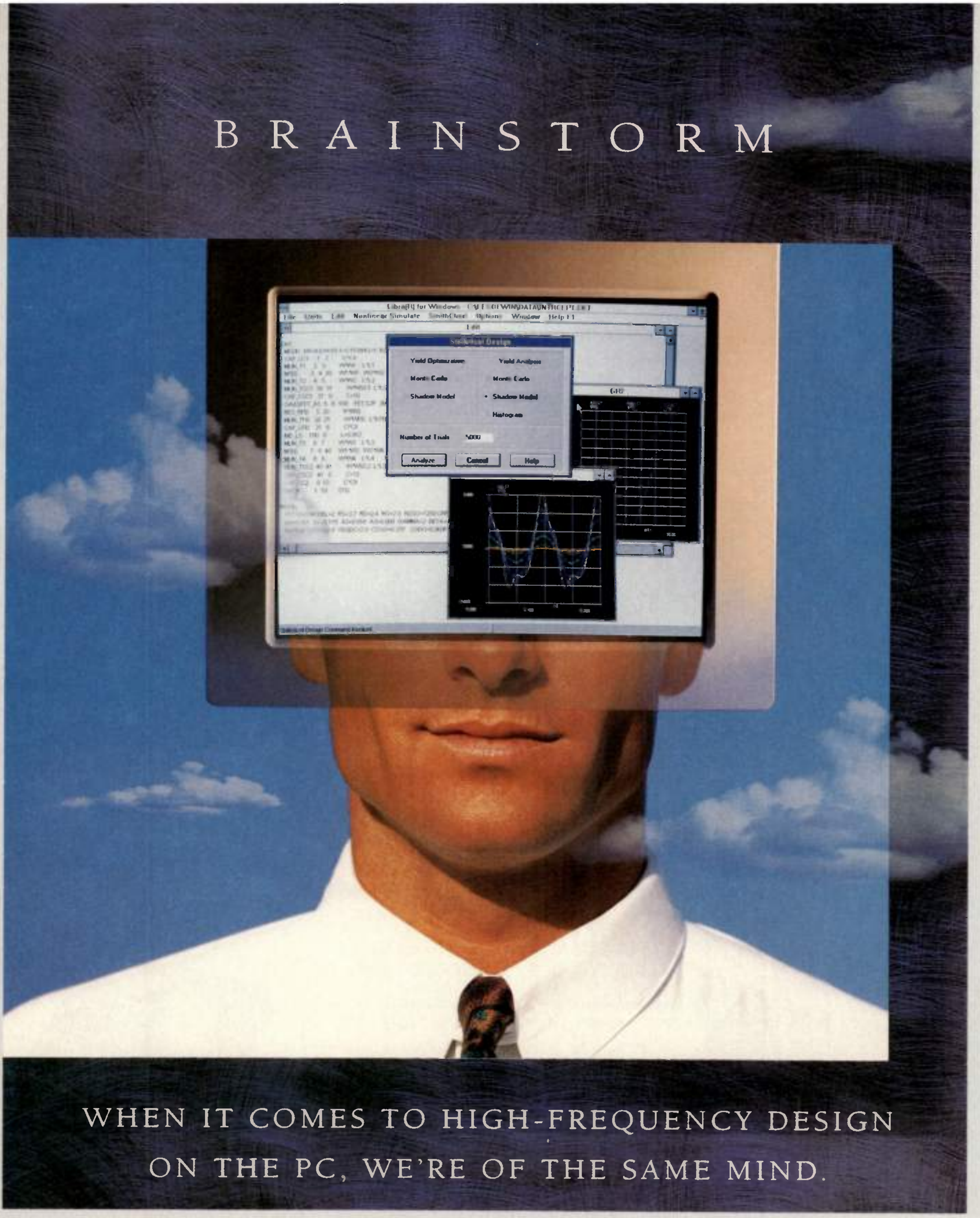

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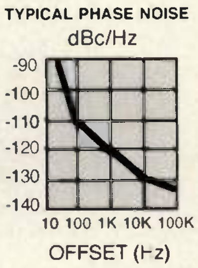
Spurs	<-45 dBc typ
Size	3.75" x 4.78" x 0.89"
Output	**Quadrature**



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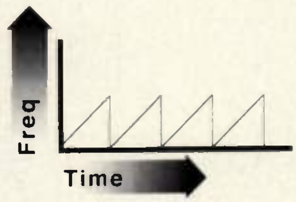
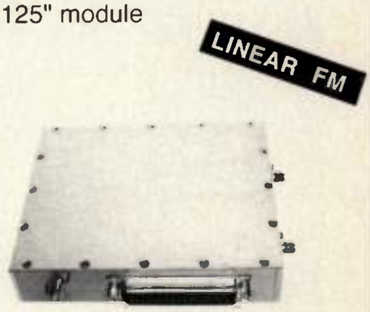
5.25" chassis or 5" x 7" x 1.125" module





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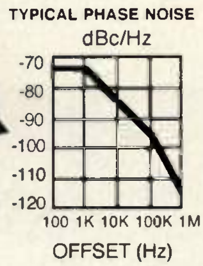

Bandwidth	**4.65 to 5.25 GHz**
Step size	2.5 MHz
Outputs	2-channels @ +17 dBm
Total power	<5 W
BIT	TTL lock indicator
Spurs	-60 dBc
Reference	10 MHz, ext (internal optional)

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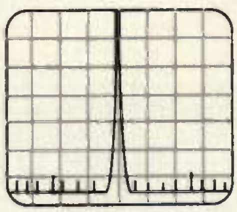

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Maximum clock	**up to 25 MHz**
Frequency control	32-bit
Switching speed	<1 µsec
Output	SINE, 1 VP-P
Phase noise	per the clock
Gating	>>100 dB ON/OFF, 1-bit toggle
Modulation	**digital amplitude, ϕ, FSK**

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