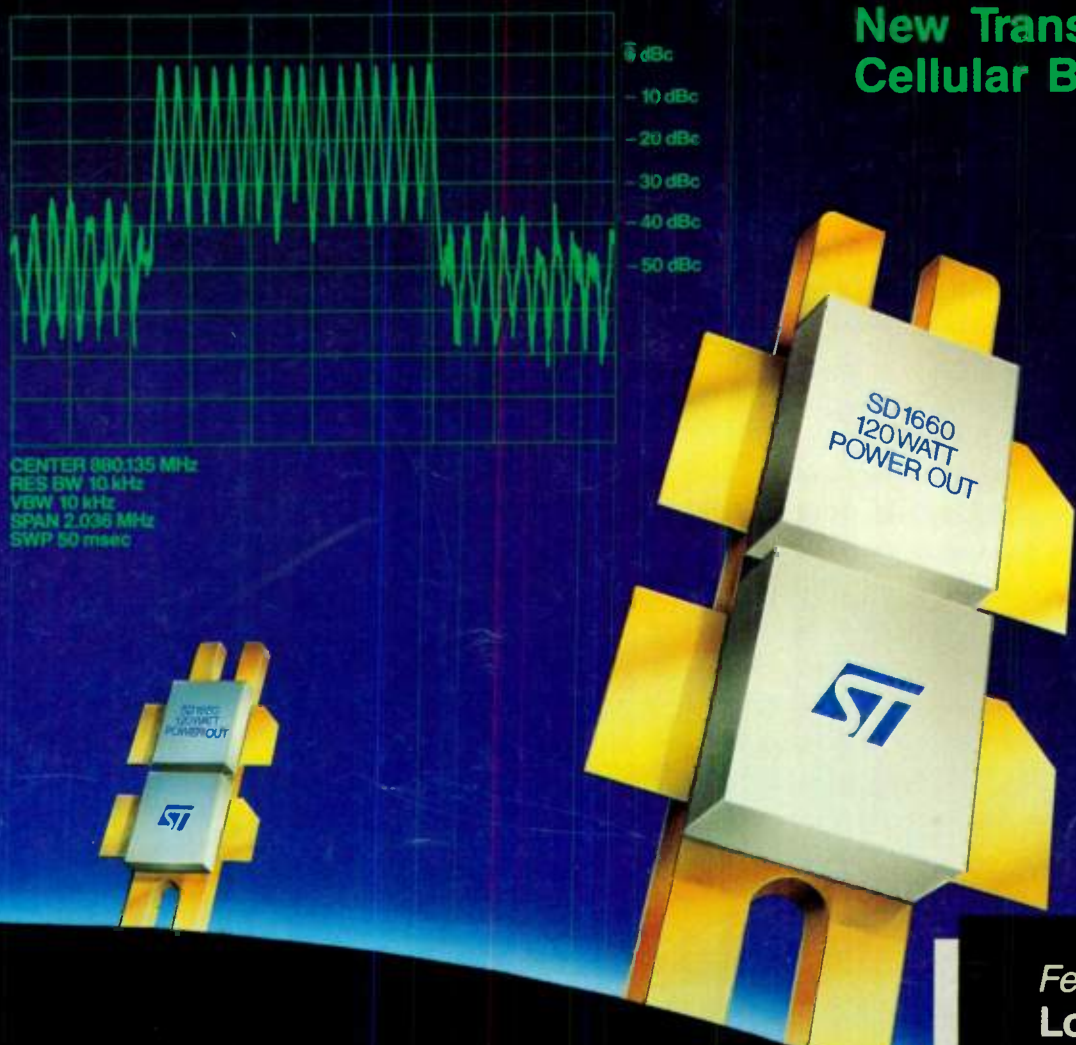
rf design

engineering principles and practices

April 1989

New Transistors Power Cellular Base Stations

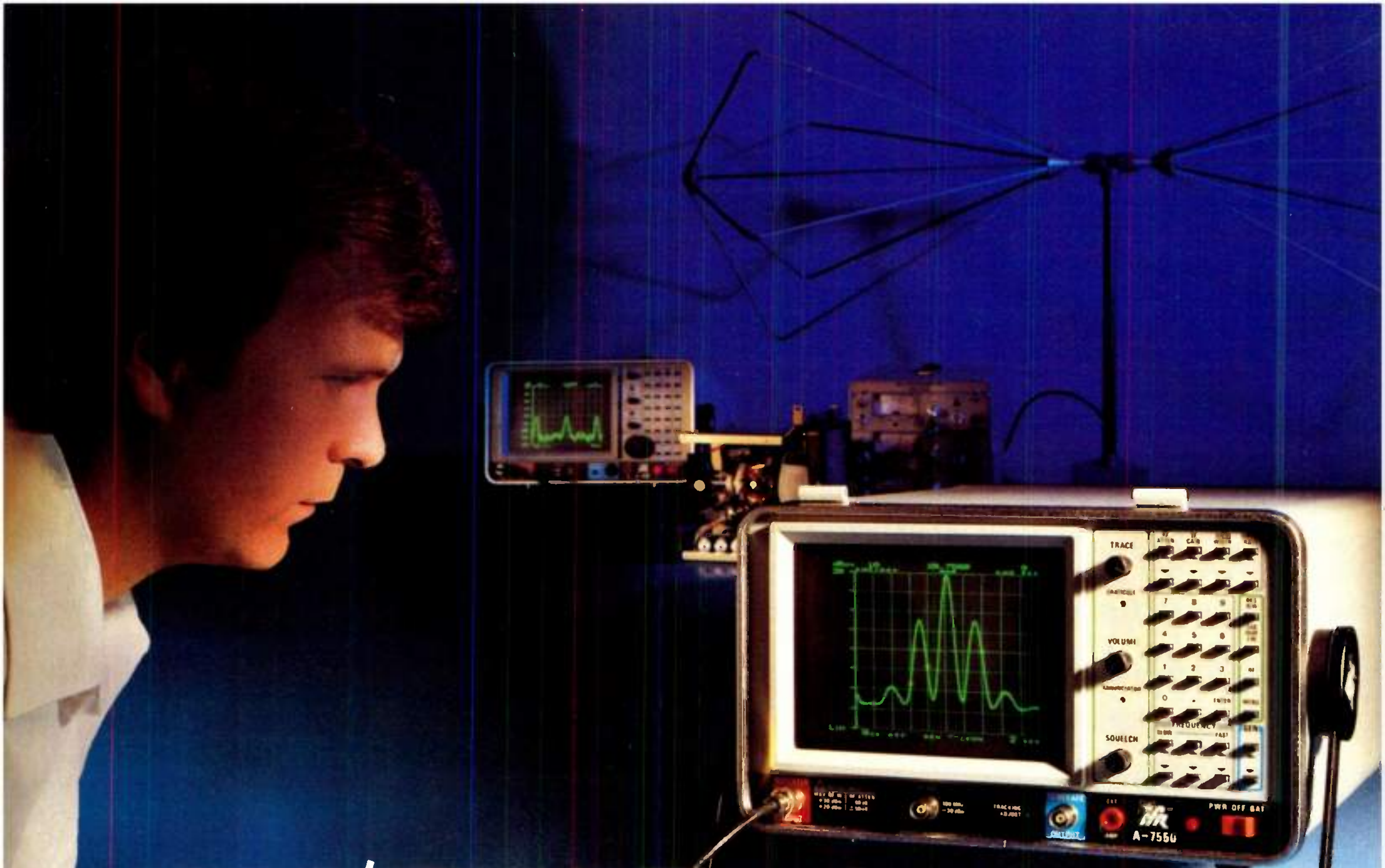


Featured Technology

Low and Medium Frequencies

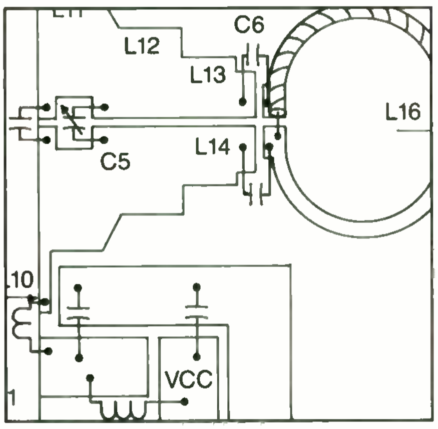
Industry Insight

RF Packaging

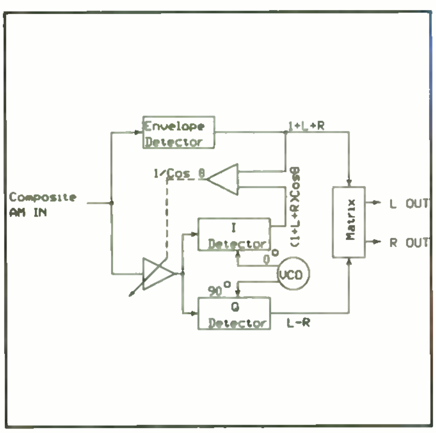

TRACE
VOLUME
SQUELCH
PWR OFF BAT

April 1989

Page 14 — RF Expo Report



Page 27 — Power Transistors



Page 54 — Phase Noise Measurement

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rf editorial

The Vitality of RF Engineering

By Gary A. Breed
Editor

In this issue is a report on RF Technology Expo 89. What the news report can't really communicate is the overall "buzz" of activity at the show. Technical discussions in the exhibit aisles, hallways, or over drinks and dinner were commonplace. The social atmosphere was fun, too, as might be expected when so many people with similar interests get together. Read the whole report on page 14.

Technical sessions featuring basic design techniques drew the biggest crowds, but the more "serious" papers generated lively discussions among some true experts in synthesizers, amplifiers, filters, oscillators and receivers. Exhibiting companies were inundated with questions on the performance and applications of their products. If you weren't there, just imagine 2450 RF engineers in one place, and I think you'll get the picture.

The importance of conferences and trade exhibitions is obvious to us. The exchange of ideas and problem-solving techniques keeps the industry vital. With this in mind, *RF Design* will be visiting or participating in several events throughout the year. Coming soon are the National Association of Broadcasters Convention, the Frequency Control Symposium, the IEEE/EMC Conference, and the IEEE MTT-S Conference and Exhibition. Look for us; we'll be delighted to see you.

Coming Attractions

Over the next three months, we will be emphasizing design techniques by frequency ranges: LF and MF this month, HF in May, and VHF-UHF in the June issue.

This issue has two articles covering the lower part of the electromagnetic spectrum. The first is a look at 90-degree constant phase shift networks, implemented primarily with active all-pass networks. Included are two means of computing the required poles: Lotus 123™ spreadsheets, and a BASIC program. The second feature is an interesting application of a consumer AM stereo decoder IC. The circuit requires precise recovery of low-level phase modulation in its primary application, which the author uses to create a simple phase noise measurement circuit.

In the next two months, you can look forward to articles on FM demodulation, intermodulation analysis, digital communications, microstrip design, high dynamic range measurements and other key topics. We hope these three "heart of RF" issues will be an interesting series, leading up to the announcement in July of this year's RF Design Awards Contest prize winners.

Gary A Breed

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Size		4 in. × 4 in. × 4 in.
Environmental	−35 to +70 C, 10,000 ft., 12 G shock. Vibration and humidity to MIL-STD 810D	
DTO		
Stability	±0.2%	±0.04%
Accuracy	±100 MHz	±1 MHz
Settling time	1 s	200 ms
SYNTHESIZER 1 MHz resolution, 10 Hz to 1 MHz modulation		
Stability	±0.03%	50 ppm
Accuracy	±50 ppm	±10 ppm

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rf design

a Cardiff publication Established 1978
Main Office:
6300 S. Syracuse Way, Suite 650
Englewood, CO 80111 • (303) 220-0600
Fax: (303) 773-9716

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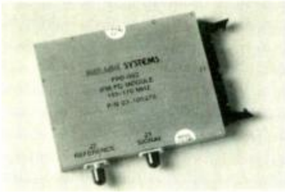
"THE TECHNICAL INNOVATION CONTINUES . . . ."



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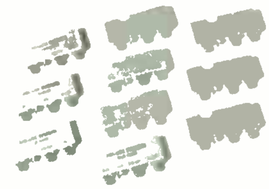
FEATURES

- Combined Instantaneous Frequency Measurement and Fast Phase Detection Circuit
- Small Size, Low Power, Surface Mount Technology



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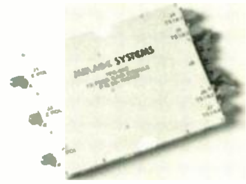

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50-500 MHz TRANSMIT FEED/GAIN UNIT

FEATURES

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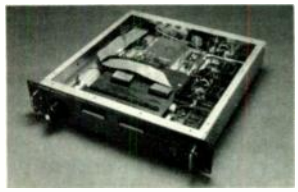
- 100 MHz to 2 GHz
- Expandable to 18 GHz



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Be part of the next generation of power solutions with Acrian as your partner.

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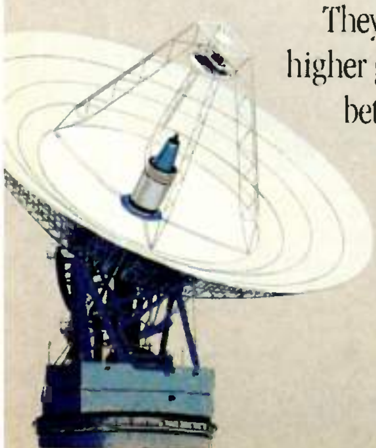
They've got 30% higher gain and 7% better efficiency than any other devices on the market.



JAM

The UDR-500 is the latest in our line-up of high-powered, wideband pulsed transistors for EW applications.

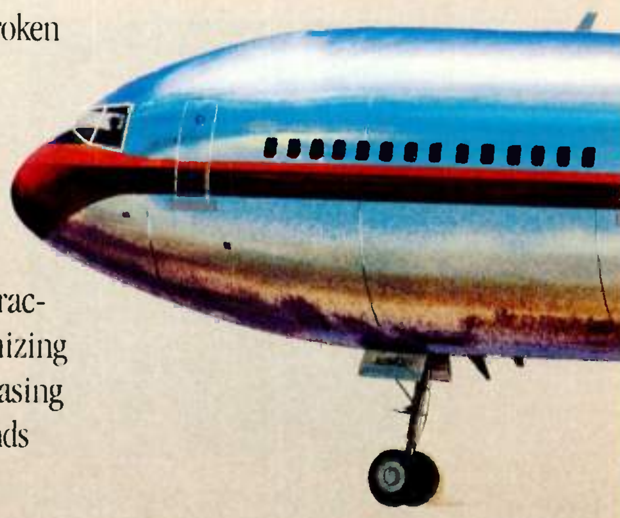
It delivers a typical 535 watts power output and a minimum of 8.5 dB gain across the UHF communications frequencies (60 μsec pulse width, 2% duty factor) making it the largest building block available for high power EW transmitters.

TRACK

Fly into the future of avionics.

Acrian transistors have broken the 150 watt power level for JTIDS, 270 watts for L-Band, and 650 watts for TCAS.

New ion implantation technology is being used to improve the performance characteristics of these parts, maximizing pulse widths/bursts and increasing duty cycles to meet the demands of next generation systems.



TION SCAN, JAM, TRACK, CALL, ACRIAN RF TRANSISTORS.

TALK

The excitement in radio communications is in wafer processing techniques. New technologies for FETs are yielding power levels up to 150 watts, and bipolar designs are pushing extra-wide bandwidths to 90 – 550 MHz at the 100 – 125 watt power levels.

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We have the winning number in cellular communications.

Improve the quality and linearity of your cellular signals with Acrian's 900 MHz line-up for base station applications. The new Class AB common emitter cellular parts will be high powered and high efficiency (60%) to ensure non-fading, uninterrupted transmission every time.

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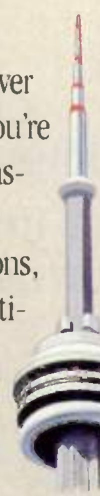

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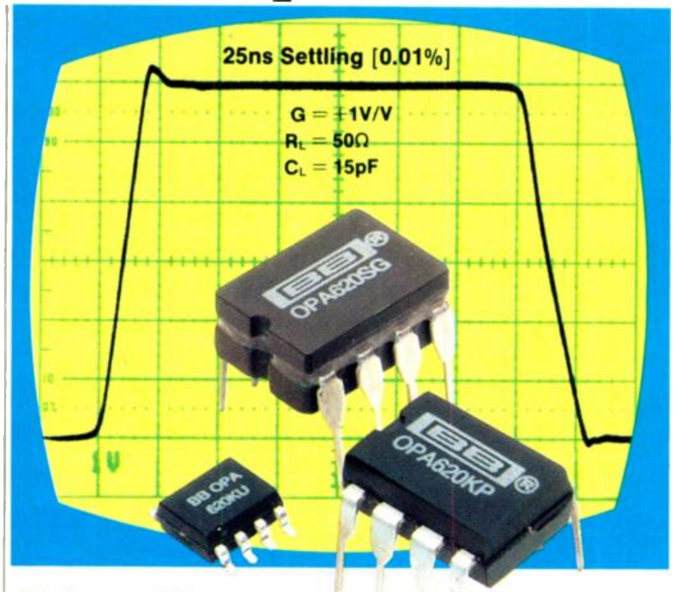
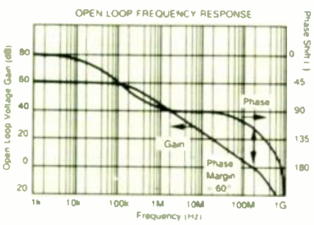
INTRODUCING

OPA620 THE COMPLETE VIDEO/RF AMP



OPA620 is a versatile new op amp for video, RF, and other high speed signal processing applications. It provides exceptional speed, accuracy, and output drive in a very easy-to-use monolithic design.



Outstanding Gain/Phase Performance

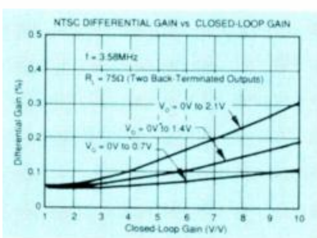
Fast Driver!

OPA620 has all the speed needed for RF/video buffering, PLL amplification, peak detecting, and signal distribution applications. Plus, its high current output lets you drive multiple cables and other heavy loads easily and economically.

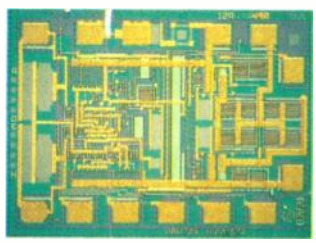
- Gain-bandwidth, 200MHz (G = +10V/V)
- Settling time, 13ns to 0.1%, 25ns to 0.01%
- Slew rate, 250V/μs
- Output drive, ±3.0V into 50Ω
- Offset voltage, ±100μV
- Drift, 8μV/°C
- Noise, 2.3nV/$\sqrt{Hz}$ @ 1MHz

Prime Time Performer

Besides high speed and drive, OPA620 also features very low differential gain (DG) and differential phase (DP) errors, important considerations for many video applications. Both DG and DP are specified at the NTSC sub-carrier frequency of 3.58MHz.



Specified Differential-Gain Error at 3.58MHz



A faster decompensated model, OPA621, is also available offering a 500V/μs slew rate and a gain-bandwidth product of 500MHz.

Classic Op Amp Design

OPA620's classic design avoids the problems of asymmetrical inputs and noisy, long settling tails associated with many current-feedback op amps. The unity-gain stable OPA620 settles fast and clean, and it's usable over a wider range of gains. Other key OPA620 features include:

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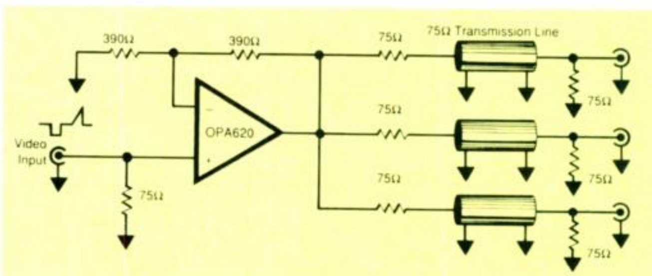

OPA620 Drives Three 75Ω Transmission Lines

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rf letters

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

Some Comments on Noise Blanking amd AM Radios

Editor:

I read with interest the article on the Sprague noise blanker IC for AM band entertainment radios (Feb 1989, *RF Design).* It demonstrates how effective pre-detection noise blanking can be in correcting what I believe to be inherent deficiencies in RF front-end design.

In a home receiver most near-field manmade noise can be eliminated by using balanced antenna coupling and a balanced (preferably loop-type) antenna. This was demonstrated by E.H. Scott back in the thirties in his famous All-Wave radio consoles. Philco also employed balanced center-tapped antenna coupling in their 1937 high fidelity models, which are extremely effective in suppressing locally generated noise. Collins Radio in their R-390/URR military receivers in the early sixties went one step further with split-stator balancing capacitors across the antenna input coils to cancel common mode interference. Obviously, a balanced antenna is required which, although practical for home reception, is not feasible for the car. Incidentally, the front-end design of most AM table radios is terribly deficient. High frequency signals can easily be received due to harmonic mixing and lack of RF selectivity.

What about the AM car radio, since supposedly no one listens to music on AM home radios anymore? The situation is no better here because car radio front-ends are designed for optimum FM antenna matching. No longer is low-capacity coax used to couple an appropriately longer whip to the receiver's antenna input. Instead, a shorter quarter-wave resonant whip and low-impedance coax are standard. Thus, AM antenna efficiency is sacrificed and the result is low sensitivity and susceptibility to ignition and alternator noise, which auto manufacturers no longer do an acceptable job of suppressing in the AM broadcast band.

Balanced antenna coupling for the home receiver and low-capacity coupling for the auto radio can significantly suppress impulse noise in the former and improve sensitivity in the latter. As far as thunderstorm static is concerned, the best assurance is a strong signal since lightning discharge noises are too long in duration to employ blanking. The use of sophisticated noise blanking is practical with today's technology only because a pound of silicon is probably cheaper than a pound of copper.

Sheldon M. Rubin
Loral Electronic Systems

A Practical Precaution From Theoretical Physics

Editor:

It is surprising how many engineers forget old Heisenberg's saying:

$$\Delta x \cdot \Delta p \geq h$$

While this is a limiting condition and is more applicable to atomic physics, the basic premise of this Uncertainty Principle is still valid in electronics — you affect the parameter you are measuring. When the nurse sticks that thermometer in you, you notice that action. By the same token, when you place a voltmeter probe on a circuit, that circuit notices it too. The thermometer may make you flinch (especially if it a cold one). The circuit also reacts — the voltage (or current or impedance) changes. It may even oscillate (if it is an amplifier). The effect may be very small, but it is there. This applies to all circuits at all frequencies (including DC). High frequency, high impedance circuits are usually more sensitive.

Fortunately, the precautions are relatively simple: use a suitable measuring device (a voltmeter with an impedance higher than that of the measured circuit), make sure that long leads to the instrument do not change that instrument's characteristics, and monitor the output to make sure that it does not change much when the instrument is applied. As simple as this sounds, these precautions are often forgotten, with the result of misleading data, which may cause unnecessary circuit changes, etc.

Andrzej B. Przedpelski
A.R.F. Products Inc.
Boulder, Colorado

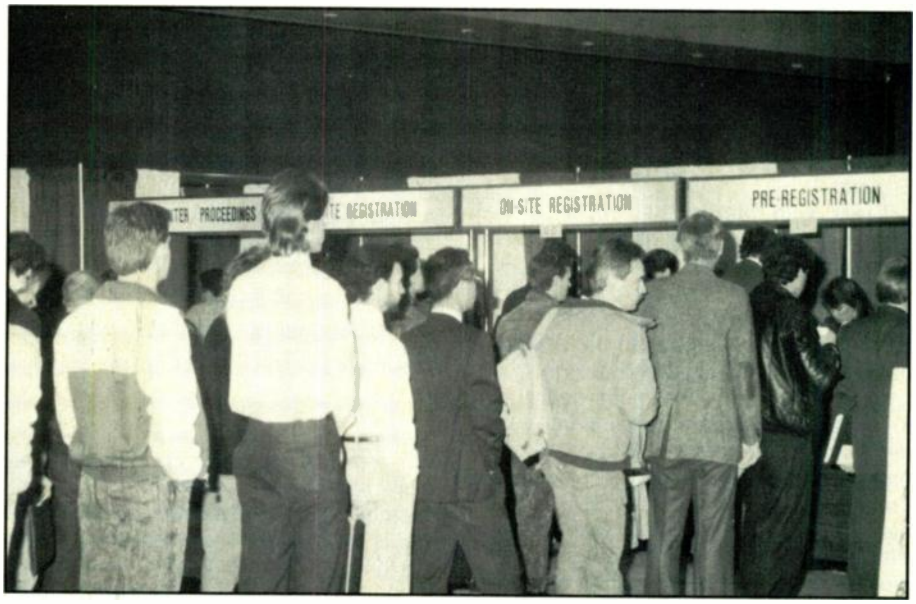
Correction

In "Design of Line Matching Networks" by Peter Martin (Feb 1989, *RF Design*), B_a was incorrectly referred to as a complex function (p. 99). Also, in Figure 3, the value of the 90 degree line, bottom left, should have been 27.4 ohms.

RF Expo 89 Has Record Attendance

2450 RF Engineers Convene in Santa Clara—With attendance up more than 30 percent over any previous RF Expo, this February's conference and exposition demonstrated that there is a high level of activity in RF technology. A very strong local turnout showed that the Silicon Valley electronics industry is a lot more than microchips and digital computers. Diversity was another characteristic evident in the crowd. Exhibiting companies who rated the show "excellent" included medical electronics firms, packaging and hardware fabricators, and primarily non-RF analog component companies, in addition to the traditional RF instrument, component and subsystem houses.

With government spending slowing, many companies seem to be looking to the commercial marketplace to sustain growth. A majority of those applications are in L-Band and lower frequencies. Many of the surviving military programs are RF systems, too, particularly in frequency-hopping communications and other electronic warfare (EW) and electronic countermeasures (ECM) applications.



2450 RF engineers crowded into the Santa Clara Convention Center for RF Technology Expo 89.

New Products Previewed

Several companies unveiled products for the first time in Santa Clara, without prior announcement. Engineers in attendance got the first look at the following products and technologies:

Acrian, Inc. chose RF Expo for a press conference and introduction of their new PSAT (Polysilicon Self-Aligned Transistor) fabrication process. Acrian claims that the new devices, beginning with UHF, L-Band and S-Band parts, have 2 dB better gain and 5-10 percentage points improvement in efficiency over current technology.

The PSAT process is derived from a method successfully used for high-speed digital integrated circuits. The key to its performance in RF devices is the ability to easily manufacture transistors with shallow emitter junctions and increased emitter periphery/base area ratios. As a result, emitter ballasting is more evenly distributed and the manufacturability increases yield and reliability.

Avantek, Inc. announced new silicon MMIC products at the show. Active mixer/amplifiers were presented, with 5 MHz to 5 GHz operating range. The IAM-81018 provides a typical RF-IF gain of 8 dB with a -5 dBm LO, and the IAM-82018 offers 15 dB gain operating from a 0 dBm LO. Also shown was the IFD-50010, an ECL-compatible frequency divider (divide-by-four) suitable for many applications up to 5 GHz. Rounding out the group is a low-noise amplifier, the INA-02170, covering DC-1 GHz with a noise figure of 2 dB and an output power of +11 dBm.

Exhibiting companies found interest high as engineers examined their products.

Many papers were delivered to standing-room-only crowds.

A patent-pending variable resolution to provide decimal frequency resolution was announced by Digital RF Solutions, whose new VR1070 direct digital synthesizer provides exactly 1.000 Hz resolution and a DC-3.3 MHz output frequency. Either an on-board reference or external precision clock (10.0 MHz) can be used.

Erbtec, which manufactures equipment and systems for medical magnetic resonance imaging (MRI) applications, brought a brand new power amplifier for the crowd to see. Intended for 1.5 Tesla systems, the MRI-20K series amplifier has a 20 kW peak linear power output, with an input level of -4 dBm. A combination of solid state and advanced vacuum tube technologies is used. Special features include extensive control and diagnostic functions, modular design, and a unique "dynamic bias" technique devised by Erbtec to compensate for temperature and device variations.

A new battery-portable microwave frequency counter from Racal-Dana was introduced at RF Expo, the Model 2101. 10 Hz to 20 GHz frequencies can be measured to 1 Hz resolution in 1 second with high sensitivity (-27 dBm at 20 GHz). Three overlapping ranges and either track or low-FM measurement modes are standard features. The unit is also capable of benchtop and ATE use as well as portable field measurements.

Sciteq Electronics announced an advance in high frequency direct digital synthesizers, an enhancement to the GaAs logic and data conversion-based ADS line. These products now achieve an output frequency of up to 300 MHz with less than 1 Hz resolution.

Microwave Technology, Inc. introduced their new solid state triode (SST) power FET line, now available in sample quantities. The SU50B12TG, a 50-watt,

Attending engineers found design solutions on the exhibit floor.

rf news Continued

RF Expo's first panel session featured Andy Przedpelski, Dan Gavin, Mike Black and Bill Egan discussing PLLs.

60-volt, 50-500 MHz device is the first of the new line. MwT also announced the success of first-generation devices designed to operate with 100-volt power supplies up to 1 GHz. The higher output impedances of these devices should make matching and combining easier and less lossy.

Courses are Popular

Two courses were presented for the first time on the west coast at RF Expo 89. Randy Rhea's Computer-Aided Filter Design class drew 65 engineers, and was extremely well-received by the students. This concentrated presentation of a single topic seemed to reach many experienced engineers who needed specific information on the design of RF filters.

At RF Expo East last November Les Besser introduced a two-part Fundamentals of RF course, with the second part devoted to deeper exploration of several important areas. Most engineers attending this class also took part in the first day's Part I session. In Santa Clara, this was again the case, with 150 engineers attending each day, a solid indication that RF continuing education is in high demand.

Technical Sessions

Many sessions were filled to standing room only, in rooms with seating for over 150. Based on the evaluation forms returned by the engineers in attendance, several speakers received exceptionally high marks in knowledge and ability to communicate. Of 40 persons presenting technical papers, 12 received a majority of "4" ratings, on a scale of 1 to 4. Six were overwhelming favorites:

Al Ward of Avantek, for his paper "UHF, L-Band and S-Band Applications of Low Noise GaAs FETs"; Mike Black of Texas Instruments, whose paper was "A Phase-Lock Loop in a Non-Cooperative Environment — A Pulsed Input"; Ed Oxner from Siliconix, who presented "Subnanosecond Switching with DMOS FETs"; Mary Holly, with FEI Microwave discussing "Design Analysis of a Space Qualified 0.8 to 1.6 GHz GaAs MESFET Low Noise Amplifier"; Donald Steinbrecher, Steinbrecher Corp., for his three-hour tutorial "Mixer Fundamentals"; and C.W. Pond of McCoy Electronics, whose paper was "A Review of Crystal Filters."

The others in the top dozen included Edward Marrone from APPCOM, Bruce Long of Piezo Crystal, Peter Bachert from Motorola, Bob Witte from Hewlett-Packard (Oscilloscopes), Barry Brown of Hewlett-Packard (Network Measurements), and Oliver Richards from Sprague Semiconductor.

These individuals, plus the other 28 speakers who contributed timely and relevant papers, were responsible for the success of the technical program. Under the direction of the Program Chairman, Dr. Tim Healy of Santa Clara University, the consensus of experienced RF Expo attendees is that this collection of papers was as good as, or better than, any previous program.

Broadcasters to Convene in Las Vegas—The 1989 National Association of Broadcasters (NAB) Convention will take place April 28-May 2, 1989 at the Las Vegas Convention Center. Over 600 exhibitors will be present on the exhibition floor, displaying equipment, products and services for the broadcasting field. An advanced television (ATV) exhibit is planned, and all proponents of ATV broadcast systems have been invited to participate in the display and demonstration of this new technology.

Of special interest to the RF engineer should be the 43rd Annual Broadcast Engineering Conference, held jointly with the exhibition. The engineering conference will offer 21 sessions and workshops with more than 100 technical papers. The program will cover a wide range of broadcast engineering topics, including technical sessions on digital audio and radio systems, AM systems engineering, radio engineering, radio production and audio processing, UHF transmission systems, and HDTV production. Engineering workshops will address RF radiation compliance, acoustics, AM antenna systems, and digital diagnostics. For further information on the convention, contact the NAB at 1771 N Street, N.W., Washington, DC 20036. Tel: (202) 429-5300

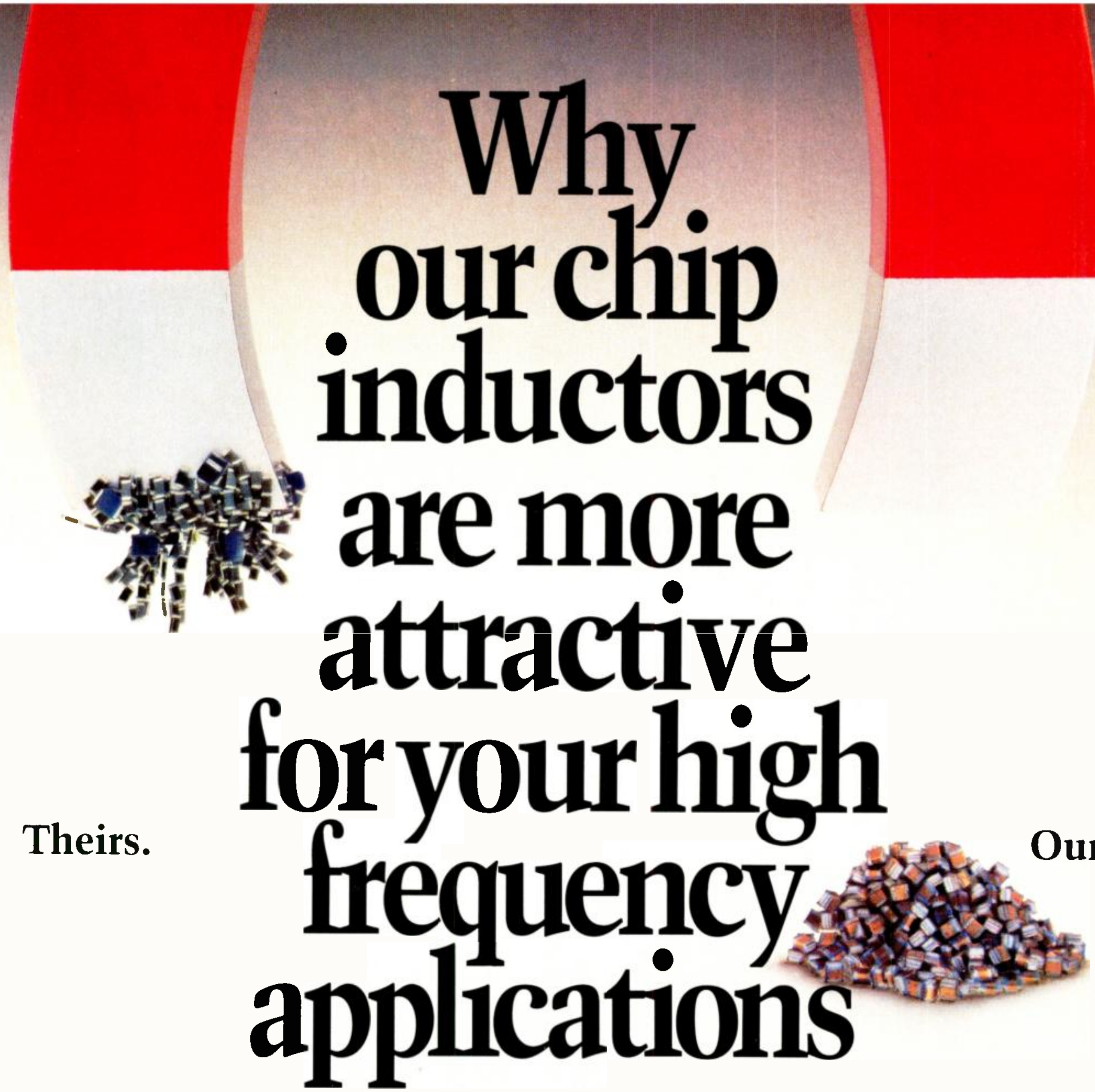
Call for Quartz Devices Papers—A call for papers has been issued for the 11th Quartz Devices Conference and Exhibition, to be held in Kansas City, Mo., August 28-31, 1989. The conference is a technical forum which addresses quartz frequency control. Papers may be tutorial or application-oriented, dealing with recent progress in design, development, processing or manufacturing control in areas represented by, but not restricted to, the following topics: properties of natural and cultured quartz; design of quartz frequency control devices; processing techniques; CAD/CAM, manufacturing and process control; measurement and test techniques and packaging; and hybrid techniques. Specific topics of interest include: Surface Mount Packaging, Improved Equivalent Circuit Models, Crystal Parameters of Importance for Oscillators and Filters, and Control of Contamination in Crystal Devices Processing. An abstract of the proposed paper, clearly describing the content, scope, organization, key points and presentation time should be submitted by April 21, 1989. Send three copies of the abstract to the Electronic Industries Association, 1722 Eye Street N.W., Washington, DC 20006.

Publication Describes Generation of EM Fields in TEM Cells—The National Institute of Standards and Technology (NIST) has pioneered in the use of transverse electromagnetic (TEM) cells for the generation of standard electromagnetic fields. Electronic equipment and/or components are inserted into TEM cells and tested for susceptibility to or emission of electromagnetic radiation. The cells are also used to calibrate portable probes for the measurement of electromagnetic fields. A new publication, *Generation of Standard Electromagnetic Fields in a TEM Cell,* documents the facilities and procedures used by NIST to generate these fields. In addition to advantages, limitations, and characteristics of TEM cells, the publication discusses setup and measurement procedures for users, uncertainties in the standard field, and statistical control of the system. The publication is available for $12 prepaid (order stock no. 003-003-02898-3) from: Superintendent of Documents, U.S. Government Printing Office, Washington, D.C. 20402.

FM-CW Radar System for Atmospheric Monitoring—Radian Corp., of Austin, Texas, has received a $1.5 million contract from the U.S. Army to build a frequency modulated-continuous wave (FM-CW) atmospheric Doppler radar. The system will be delivered to White Sands Missile Range in 1989, following extensive testing in Boulder, Colo., at the National Oceanic and Atmospheric Administration's (NOAA) Boulder Atmospheric Observatory. The radar system will run continuously for five years at White Sands, providing real-time and cumulative atmospheric information necessary for Strategic Defense Initiative directed energy research.

The FM-CW atmospheric Doppler radar is a synthesis of two different radar techniques: high resolution continuous wave and Doppler technology. The end product is a powerful radar capable of operating in all types of weather and at short range as well as long. This radar can detect invisible clear-air turbulence, wind speed and direction, and all forms of hydrometeors. It will also be the highest resolution radar system of this type available, capable of resolving less than a meter in range.

Due to project and environmental specifications required by the Army, the FM-CW atmospheric radar system will be portable. Two parabolic antennas will be mounted on a trailer and connected to a van which will house the computer equipment for the radar system. Data will be available on site as well as remotely accessed.



The point of this little demonstration is that Coilcraft surface mount inductors are made of ceramic. A decidedly non-magnetic material.

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So next time you're selecting surface mount inductors, forget the ferrite and stick with Coilcraft ceramic chips.

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INFO/CARD 18

RF Packaging — To BeO or Not to BeO

By Katie McCormick
Assistant Editor

The demand for improved performance and reduced cost represents a real challenge to today's designer of electrical devices. In the RF world, the pursuit of increased power, higher speeds and greater densities translates ultimately into a problem of heat. Not surprisingly, one subject receiving increasing attention from designers and manufacturers is the issue of packaging. In fact, in terms of performance as well as cost, packaging is now viewed as a critical limitation. "Miniaturization is great from a performance point of view," says Steve Lockard, market development supervisor for Rogers Corp., "but it creates problems with thermal management."

At present, there is only one packaging material in widespread use which addresses the need for the efficient dissipation of heat demanded by high power RF applications. That substance is beryllia (beryllium oxide, or BeO) — a ceramic material whose properties of high thermal conductivity and low thermal resistance, combined with an ability to be easily metallized, make it an ideal substrate material for use in RF power transistor packages. Ideal, that is, at first glance. In reality, the subject at the forefront of any discussion of RF packaging is the search for a replacement for beryllia. "The big issue for packaging is the use of BeO," observes Bert Berson, president of Berson and Associates, a strategic planning firm with strong involvement in packaging.

There are several problems associated with the use of BeO. The first, and most urgent, is its toxicity — exposure to BeO dust can cause berylliosis, a disease similar to asbestosis. Use of BeO in the manufacturing process requires extra precautions, and thus extra cost. Bob Shaw, sales manager for Cabot Ceramics' Microwave Division, notes that another cost factor is the fact that BeO is supplied by only one vendor (Brush Wellman, Cleveland, Ohio). "They control price and delivery to all packaging vendors," he explains. The main problem seems to be finding a material which can produce the same level of performance as beryllium oxide for higher power applications. "We use all the thermal conductivity we can get out of BeO," stresses Tom Rice, senior manufacturing engineer with Motorola's RF Optoelectronics Products Division. Any replacement material must be at least as good, if not better, he adds.

Several materials are being evaluated as potential substitutes for BeO. "The immediate trend," according to Allan Buck, sales manager for Kyocera America's Metallized Products Division, "is to go to aluminum nitride." Generally believed to be the most promising candidate, aluminum nitride has the thermal conductivity and expansion characteristics needed, and will likely be supplied from several sources rather than only one. "The problem with aluminum nitride as a replacement for beryllia is in finding a good metallization," notes Bob Shaw.

Another material mentioned as a possible replacement for BeO has been boron nitride, but there are problems. "It doesn't have the thermal conductivity that we need, it's cost-prohibitive, and not as readily manufactured here in the United States," Rice points out. Diamond may be a good replacement but research into its applications is not as far along as for aluminum nitride, and the cost factor is significant.

Several companies are currently involved in investigating aluminum nitride's potential, and estimates about its eventual use as a replacement for BeO are optimistic. "If in fact the material does turn out to be O.K.," comments Rice, "I see limited production some time next year." Dennis Duff, vice-president, sales and marketing for Keramount Corp., an aluminum nitride powder manufacturer, also anticipates production-level use of the substance by next year. His long-term prediction — "A significant amount of beryllia consumption should be replaced by aluminum nitride."

As mentioned, a motivating factor in the search for a BeO substitute is the desire to reduce cost. One explanation for this, according to Cabot's Bob Shaw, is the shift he is observing from military to commercial applications among packaging customers. "They're being forced to be cost-sensitive to be competitive," he says. In addition to rising materials costs, Shaw points out that the cost of the package as a percentage of total device cost is on the rise. He sees this as a key reason behind the increased attention being paid in the industry to packaging. Efforts aimed at reducing costs include the move to offshore production in order to reduce labor rates, a greater focus on increasing productivity and yields, and the incorporation of surface mount technology and automation into the production process.

Bert Berson agrees that cost is of prime concern in the packaging industry right now. One step toward a solution which he advocates is the implementation of some packaging standards. Berson notes the enormous proliferation of different kinds of packages, along with the time and research involved in getting new package designs. These are some of the problems being addressed at the Electronic Packaging Center at Washington University, St. Louis, Mo., established a year ago by Berson and Fred Rosenbaum, the center's director.

As far as overall market trends go, GE Ceramics' marketing and sales manager Normand Allard anticipates that the ceramic packaging market will continue its growth. "The market," he says, "is definitely growing; I guess my biggest fear is whether it will stay in the United States." He advises that American firms "ought to start supporting their U.S. suppliers." Allan Buck of Kyocera sees growth opportunities in European commercial markets. In addition, he predicts, "The long-term trend is to go to modules."

What impact do these concerns have on the design process? "We're seeing a lot more involvement and concern over packaging in the last year from our customers," states Bob Shaw. "Our customers, the device manufacturers, are putting that pressure to reduce cost on their designers," Allan Buck reports. These trends should continue as new technologies offer the design engineer more freedom and greater performance, in the midst of increasing emphasis on competitiveness and cost. rf

rf calendar

April 25-27, 1989
IEEE Instrumentation and Measurement Technology Conference
Key Bridge Marriott Hotel, Washington, DC
Information: Robert Myers, 1700 Westwood Boulevard, Suite 101, Los Angeles, CA 90024. Tel: (213) 475-4571

April 26-28, 1989
Aerospace and Defense 89
Santa Clara Convention Center, Santa Clara, CA
Information: Chuck Jungi, American Electronics Association, 5201 Great American Parkway, Santa Clara, CA 95054. Tel: (408) 987-4202

April 29-May 2, 1989
National Association of Broadcasters 67th Annual Convention and 43rd Annual Broadcast Engineering Conference
Las Vegas Convention Center, Las Vegas, NV
Information: National Association of Broadcasters, Conventions and Meetings, 1771 N Street, N.W., Washington, DC 20036. Tel: (202) 429-5300

May 17, 1989
Emerging Microwave Technologies Symposium and Show
Crest Hollow Country Club, Woodbury, NY
Information: Pari Boloori, Narda Microwave Corporation, 435 Moreland Road, Hauppauge, NY 11788. Tel: (516) 231-1700 ext. 437

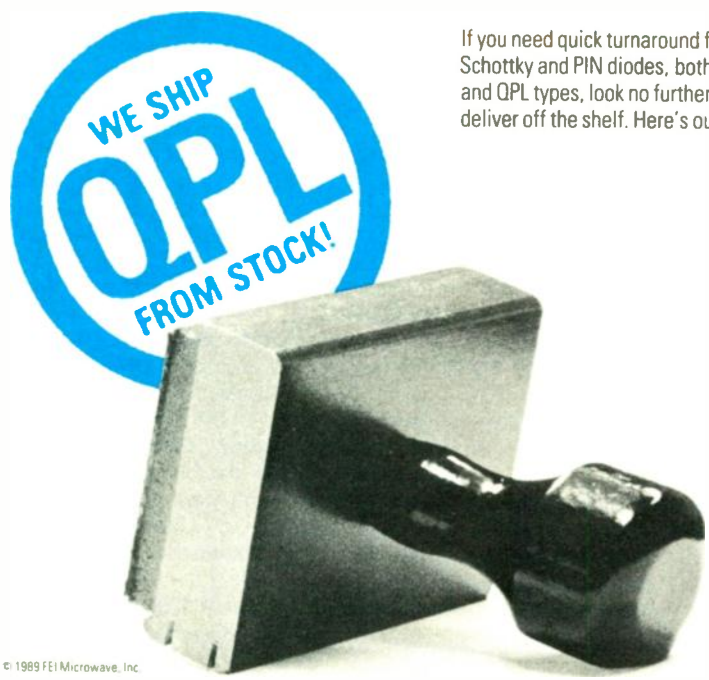
May 22-24, 1989
39th Electronic Components Conference
The Westin Hotels, Houston, TX
Information: EIA, 1722 Eye Street, N.W., Washington, DC 20006. Tel: (202) 457-4930

May 23-25, 1989
IEEE 1989 National EMC Symposium
Radisson Hotel, Denver, CO
Information: Jon Tary, Tri-State, 12076 Grant Street, Denver, CO 80233. Tel: (303) 452-6111

May 31-June 2, 1989
43rd Annual Frequency Control Symposium
Denver Marriot Hotel — City Center, Denver, CO
Information: Michael Mirachi, Synergistic Management Inc., 3100 Route 138, Wall Township, NJ 07719. Tel: (201) 280-2022

June 5-7, 1989
NBS/Boulder Time and Frequency Seminar
Boulder, CO
Information: David Allan, NIST, 325 Broadway, M/S 576, Boulder, CO 80303. Tel: (303) 497-5637

June 15-16, 1989
33rd Annual ARFTG Conference
Long Beach Convention Center, Long Beach, CA
Information: ARFTG, P.O. Box 2182, Longmont, CO 80502.



rf courses

The George Washington University

Frequency-Hopping Signals and Systems
May 8-10, 1989, Washington, DC
Monopulse Radar Principles and Techniques
May 8-11, 1989, Washington, DC
Radar ECM and ECCM Systems
May 8-12, 1989, Washington, DC
Radar Systems and Technology
May 8-12, 1989, Washington, DC

Information: Misael Rodriguez, Continuing Engineering Education, George Washington University, Washington, DC 20052. Tel: (800) 424-9773; (202) 994-6106

Compliance Engineering

EMI
April 25, 1989, Boston, MA
June 20, 1989, Chicago, IL
Safety
April 26, 1989, Boston, MA
June 21, 1989, Chicago, IL
August 23, 1989, San Jose, CA
ESD
April 27, 1989, Boston, MA
June 22, 1989, Chicago, IL
August 24, 1989, San Jose, CA
Telecom
April 28, 1989, Boston, MA
June 23, 1989, Chicago, IL
August 25, 1989, san Jose, CA

Information: Compliance Engineering, 629 Massachusetts Avenue, Boxboro, MA 01719. Tel: (508) 264-4208.

EEsof Inc.

MMIC Design Workstation (MMIC)
April 24-28, 1989, Westlake Village, CA
System Design (OmniSys)
May 11-12, 1989, Westlake Village, CA
CAD for Nonlinear Microwave Circuits (mwSpice)
May 18-19, 1989, Westlake Village, CA

Information: Sande Scoredos, Training Coordinator, EEsof Inc., 5795 Lindero Canyon Road, Westlake Village, CA 91362. Tel: (818) 991-7530, ext. 197

EMC Services

EMI Control for Switching Supplies
May 1-5, 1989, San Diego, CA
July 24-28, 1989, San Francisco, CA
Filter Design for Switching Supplies
May 1-5, 1989, San Diego, CA
July 24-28, 1989, San Francisco, CA

Information: Sonya Nave, EMC Services, 11833 93rd Avenue North, Seminole, FL 34642. Tel: (813) 397-5854

rf courses *Continued*

Henry Ott Consultants

Workbench EMC Measurements and Troubleshooting
May 1-2, 1989, East Hanover, NJ
Electrostatic Discharge Protection and Control
May 3-4, 1989, East Hanover, NJ

Information: HOC, 48 Baker Road, Livingston, NJ 07039. Tel: (201) 386-5622

Integrated Computer Systems

C Programming Hands-On Workshop
April 18-21, 1989, Boston, MA
April 25-28, 1989, San Diego, CA
May 2-5, 1989, Indianapolis, IN
Fiber Optic Communication Systems
April 25-28, 1989, Los Angeles, CA
May 9-12, 1989, Washington, DC
June 6-9, 1989, Boston, MA
Digital Signal Processing: Techniques and Applications
May 2-5, 1989, Los Angeles, CA
May 16-19, 1989, San Francisco, CA
May 23-26, 1989, Washington, DC

Information: John Valenti, Integrated Computer Systems, 5800 Hannum Avenue, P.O. Box 3614, Culver City, CA 90231-3614. Tel: (800) 421-8166; (213) 417-8888

Interference Control Technologies, Inc.

EMC Design and Measurement
May 1-5, 1989, Palo Alto, CA
Practical EMI Fixes
May 1-5, 1989, Chicago, IL
Grounding and Shielding
May 9-12, 1989, Orlando, FL

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

R & B Enterprises

Real Life Solutions to EMI Problems
May 2-4, 1989, Washington, DC

Information: Registrar, R & B Enterprises, 20 Clipper Road, West Conshohocken, PA 19428. Tel: (215) 825-1966

UCLA Extension

Microwave Circuit Design I: Linear Circuits
June 19-23, Los Angeles, CA
Microwave Circuit Design ii: Non-linear Circuits
June 26-30, Los Angeles, CA

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

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INFO/CARD 24

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INFO/CARD 25

New Transistors for the Next Generation of Cellular Base Stations

By Serge Juhel, John Walsh, and William Imhauser
SGS-Thomson Microelectronics

The next generation of cellular telephone systems will require cell sites with increased power output and linearity. RF power amplifiers will need the capability for 100-150 watts PEP (16 channels), and have total intermodulation products below -60 dB (approximately -30 dB in the amplifier, and the rest obtained by signal processing techniques). Until recently, this high power could only be achieved by paralleling several devices, with increased power consumption due to combining inefficiencies and large size. This article describes new high-power devices for this demanding application, and the description of a 120-watt amplifier circuit.

Two new high power common emitter devices especially designed and processed for class AB cellular base station applications have been introduced by SGS-Thomson: the SD1660,

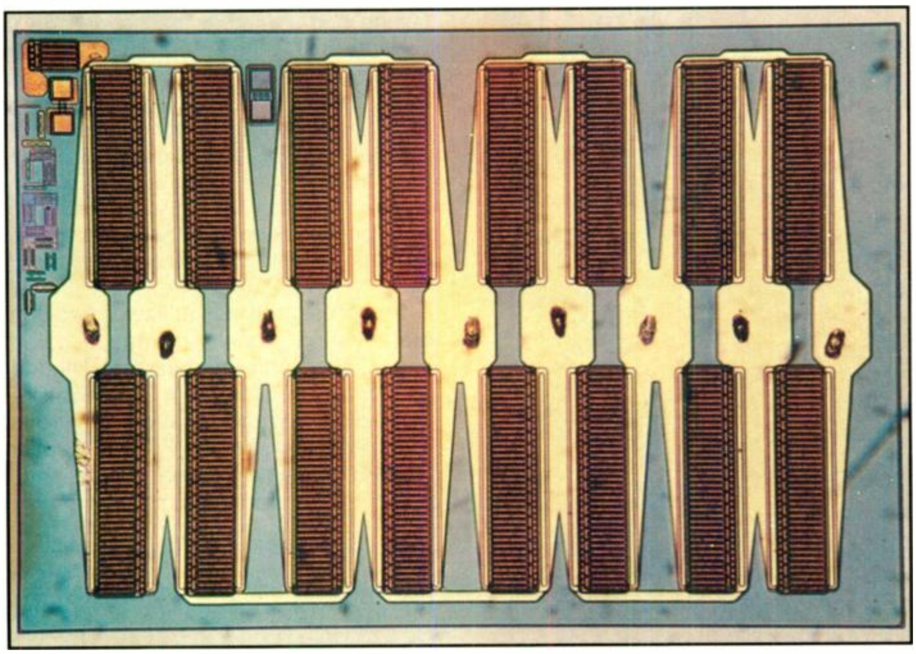

Figure 1. Interdigitated die structure used in the SD1660 and SD 1680.

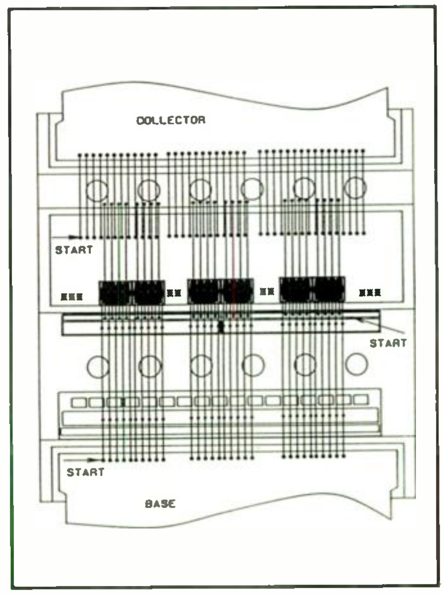

Figure 2. Package layout using plated-through holes for emitter connection.

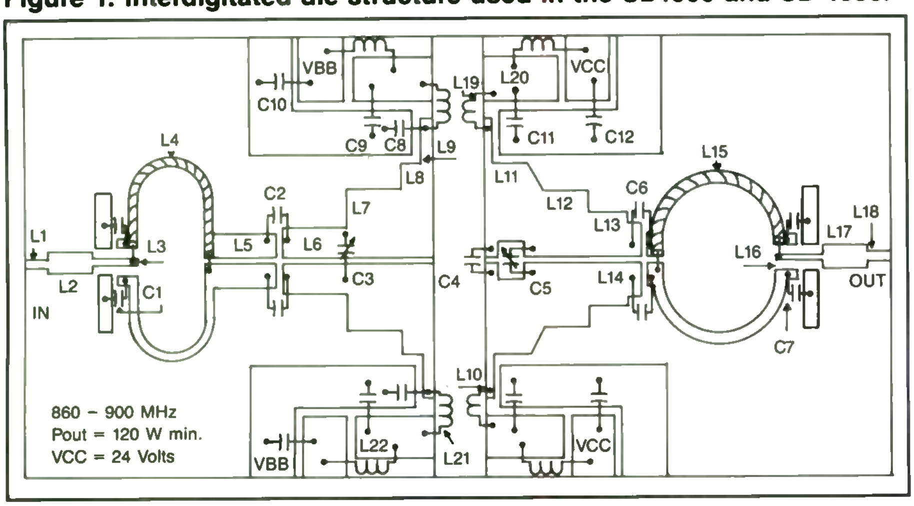

Figure 3. 120-watt amplifier circuit diagram and p.c. layout.

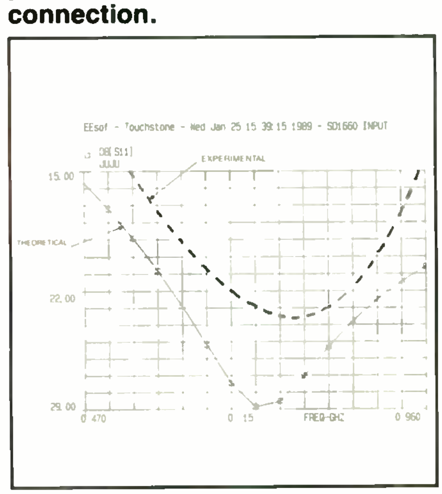

Figure 4. Theoretical vs. measured performance of balun.

a 120 W PEP, 24 V device for 860-900 MHz with 6.0 dB gain and intermodulation distortion (IMD) products ⩽ −32 dB; and the SD1680, optimized for 915-960 MHz with 100 W PEP, 6.0 dB minimum gain, and IMD ⩽ −32 dB. Both are balanced, push-pull devices.

These devices are the result of a two-year development program to attain a 150 W (minimum) class AB linear transistor for the UHF television band, 470-860 MHz. The SD1492, developed through this program, has been in production for a year, and is now used in television transmitter designs up to 30 kW. The starting material, emitter ballast levels, and base diffusion of this die have been optimized to create the SD1660 and SD1680 for higher-frequency cellular base station applications. Broadband application circuits have been developed for these transistors and accompanying pre-driver devices.

The Transistor Die

Interdigitated structures, although difficult to process, offer the best performance for linear applications. Excellent emitter periphery utilization (related to saturated power) allows the use of very high levels of emitter ballast, achieving maximum linearity for a given power gain. The die (Figure 1) for these new cellular devices has three forms of ballasting. The primary ballast medium is a diffused resistor which has a "pinch off" effect at high current levels. Also, a polysilicon resistor offers finger ballasting and some site ballasting due to contact resistance at the interface with the single crystal. This contact resistance presents a high impedance to localized hot spotting tendencies, which are severe in DC-biased class AB operation. Another benefit of the polysilicon-covered emitter is relative immunity from emitter-base leakage, a significant problem in yield for interdigitated structures.

The present device utilizes a two-step metallization process to reduce output capacitance and improve metal feeder efficiency by allowing a much thicker second metallization of about 3 microns. The geometry has a 7 micron pitch (emitter plus base contact repeats every

f MHz	Z_{IN} Base to Base	Z_{LOAD} Collector to Collector
850	6.8 + j3.5	4.4 +j1.6
860	6.5 + j3	4.3 + j2
880	5.9 + j2.1	4.2 + j2.8
900	5.2 + j1.6	4.3 +j3.7

Z_{LOAD} = Optimum load impedance into which the transistor operates at the specified power out, voltage and frequency.
860-900 MHz Broadband Amplifier:
P_{OUT} = 120 Watts
V_{CC} = 24 Volts
Class AB

Table 1. SD1660 input/output impedances.

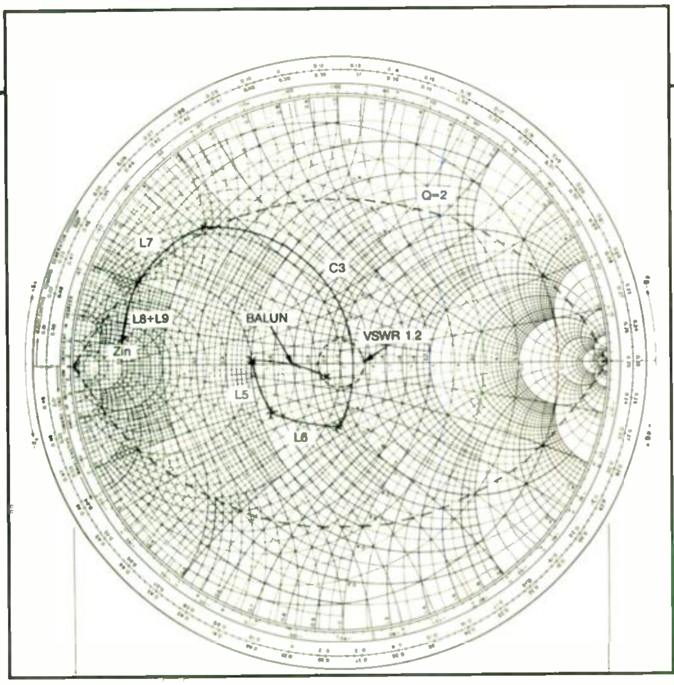

Figure 6. Input performance at 900 MHz.

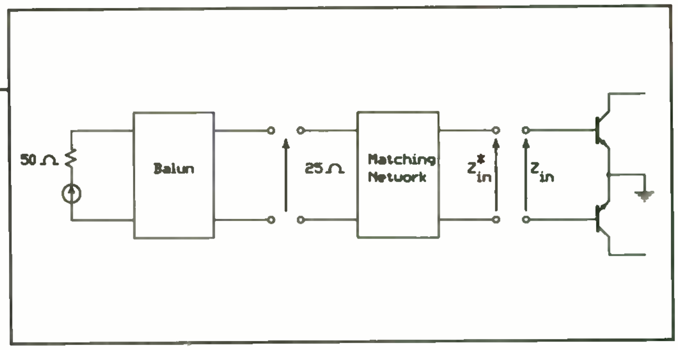

Figure 5. Input circuit configuration.

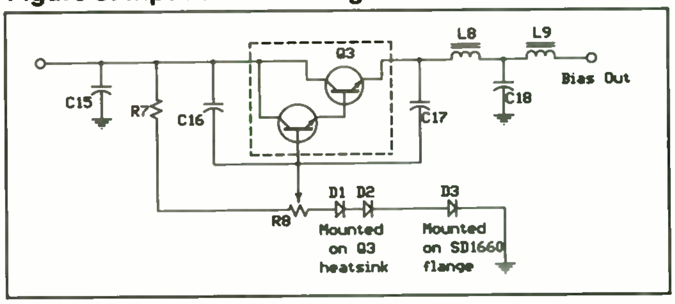

Figure 7. Bias network.

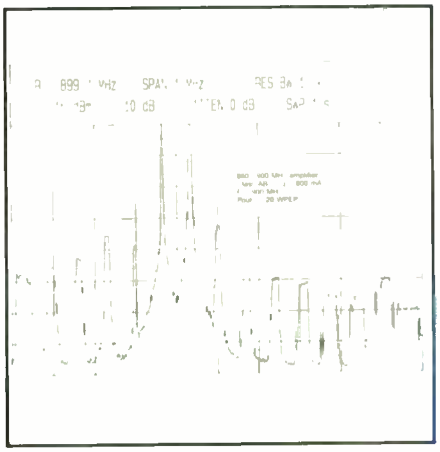

Figure 8. Two-tone IMD test results.

7 microns). The emitter is 0.8 microns wide and contacts are submicron to allow full coverage by a 2 micron metal finger.

The device performance can be attributed to a basic 3 GHz geometry which utilizes shallow ion implantation for all diffusions. Thus, an S-band geometry and diffusion is combined with a very high level of ballasting to achieve a very linear device with acceptable gain at 900 MHz. A single die produces 20 watts class AB. Die are combined in a package, along with an input matching network to optimize performance over a given bandwidth.

The SD1660 operates in a common emitter configuration, and must have low emitter inductance. To reduce the negative feedback and accompanying reduction in gain caused by emitter inductance, SGS-Thomson uses a package with plated-through ground holes (Figure 2).

120 W Amplifier Circuit

Figure 3 shows the layout of an amplifier for 860-900 MHz using the SD1660 to obtain 120 watts class AB output. Input and output impedances have been measured from base to base and collector to collector (Table 1). The circuit is built on a 30-mil PTFE glass p.c. board, $\varepsilon_r = 2.55$.

Since the SD1660 is a push-pull device, the unbalanced input has to be transformed to the balanced base input. A balun transformer is used to simulta-

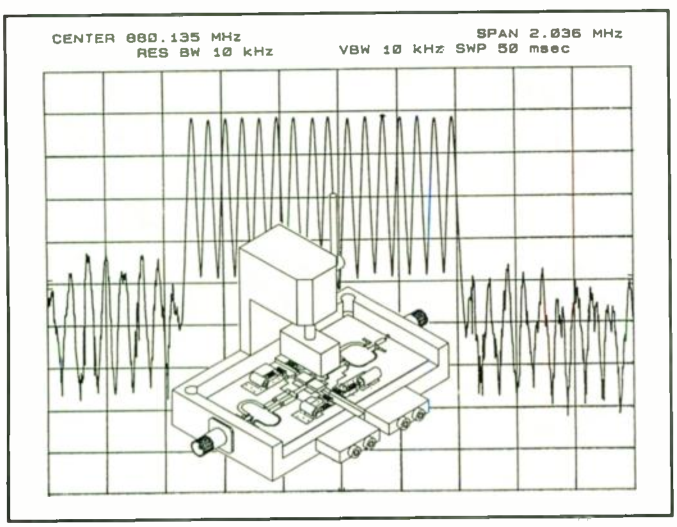

Figure 9. 16-tone test results, showing greater than -30 dBc IMD products.

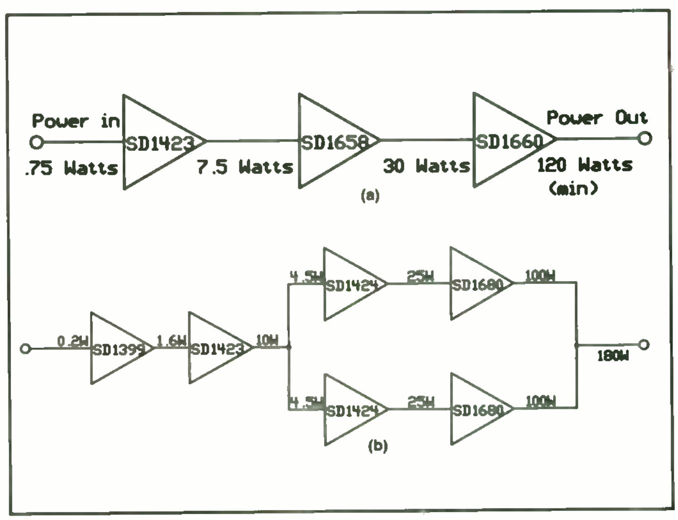

Figure 10. Device lineups for 860-900 MHz (a) and 915-960 MHz (b).

neously reduce impedance, maintain large bandwidth, and provide a balanced output. Figure 4 shows the performance of the 50 ohm to 2x12.5 ohm balun circuit used in the amplifier.

Next, consider the schematic in Figure 5. It can been seen that the matching network must match 25 ohms to the device input impedance. The network used consists of lowpass sections. C3 is adjusted for optimum VSWR in the band. Figure 6 demonstrates input performance at 900 MHz.

The output matching is accomplished in the same fashion as the input, with two distinct sections: a matching network transforming the output impedance to 25 ohms (balanced), and a balun to transform the balanced 25-ohms to the 50-ohm unbalanced output.

In class AB, the transistors are biased just at the start of conduction, about 400 mA per side in the case of the SD1660. To maintain a stable quiescent current, a diode (D3 in Figure 7) is used to compensate for the variation in temperature (-2 mV/°C) of the base-emitter junction of the RF device. This diode should be mounted as close as possible to the device flange, to get the best thermal tracking. D1 and D2 are compensation for the darlington pass transistor Q3. Compensation of quiescent current is a critical parameter when considering intermodulation distortion (IMD). Also, the bias network must be able to supply sufficient current for proper transistor operation.

Linearity

The most convenient method of measuring the linearity of an amplifier is the multicarrier/IMD product test, with the amplifier operating at rated PEP output level. In this case the power read by a power meter is:

$$P_{average} = PEP/n$$

where n = number of carriers

The level of IMD products generated in such a test is a measure of linearity. Figure 8 shows the SD1660 amplifier two-tone test results, while Figure 9 is the same amplifier under a 16-tone test. The IMD is greater than -30 dBc, and with signal processing techniques, -60 dBc can be achieved.

Using the device lineup shown in Figure 10(a), a 22 dB gain amplifier subsystem can be constructed for the 860-900 MHz band. For the Pan European Radiophone system, which operates in the 915-960 MHz band, the devices in Figure 10(b) can be used to obtain 180 W with 29.5 dB gain.

Readers interested in more information about these transistors can contact the authors at SGS-Thomson Microelectronics, 211 Commerce Drive, Montgomeryville, PA 18936-1002, or by telephone at (215) 362-8500. Information can also be received by circling INFO/CARD 225.

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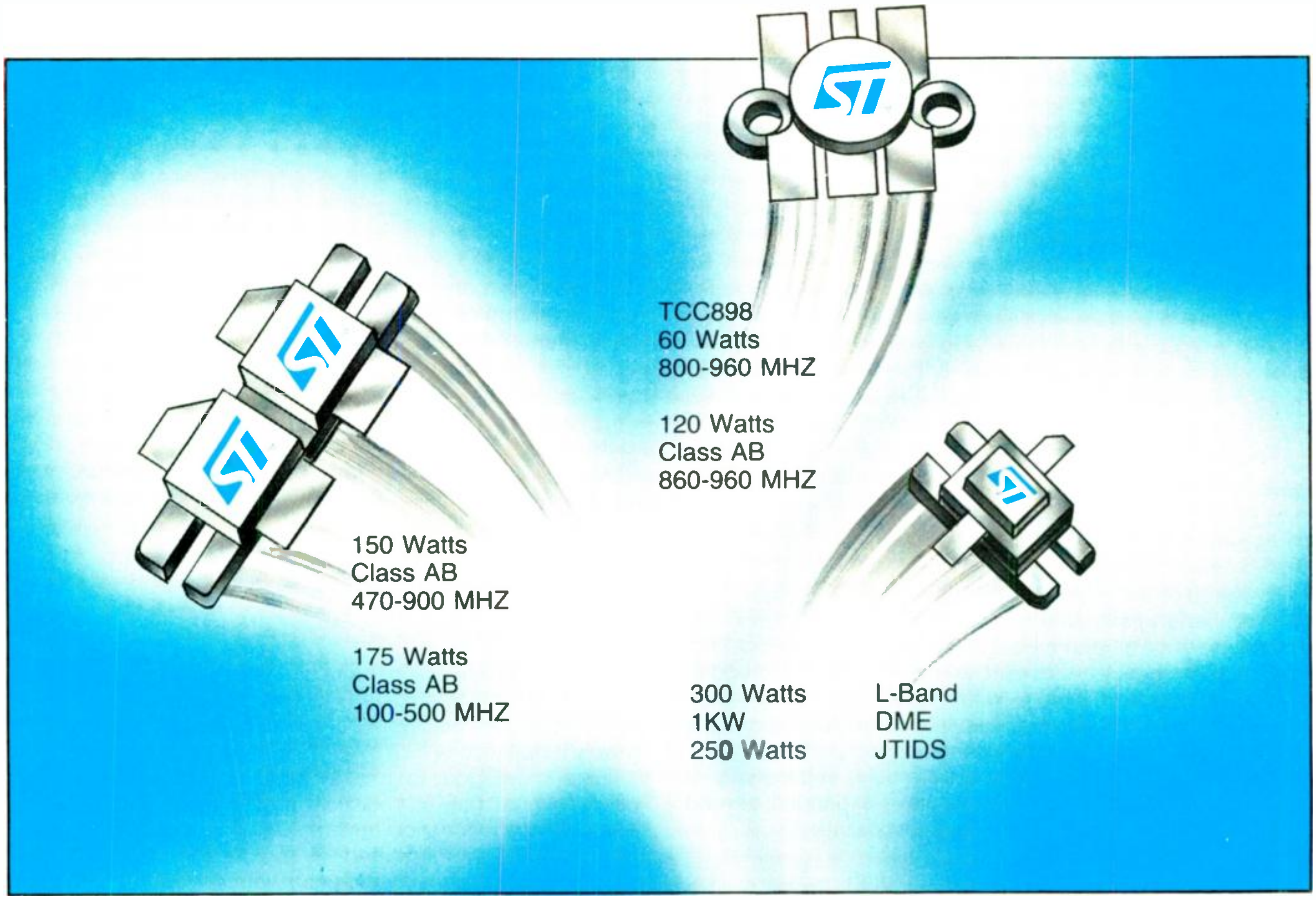

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INFO/CARD 28

Design of Constant Phase Difference Networks

By Thomas A. Keely
Honeywell Inc.
Military Avionics Division

A standard topic of college electronic communications courses is the synthesis of single sideband signals. Typically, textbooks discuss two methods of sideband synthesis: the filter method, which selectively filters one of two sidebands produced by suppressed carrier modulation, and the phase shift method, also known as the Hilbert Transform (1), where the carrier and the baseband modulating signal are mixed in phase quadrature. This article describes the design of wideband phase difference networks for implementation of the phase shift method, meeting specified phase shift criteria, phase shift error bounds and bandwidth criteria.

A limitation of the phase shift method is the requirement for a wideband phase shift or phase difference network that would keep the baseband modulating signal in 90° phase shift over a wide range of frequency. Darlington's discussion (2) of realizing a constant phase difference over a specified bandwidth treats the theoretical aspects of the subject, but it's easy to get lost in the elliptic functions.

In the time since Darlington's 1950 article, applications of wideband phase difference networks have expanded beyond single sideband synthesis, including the following:

- Signal generators with outputs in quadrature.
- I and Q receivers which require local oscillator signals in phase quadrature.
- Digital filter phase difference networks requiring s-plane poles and zeros which can be transformed into the time sampled z-plane.

This article describes one important method used to realize a constant phase difference, in which two allpass networks are employed with the difference between the two phase transfer functions producing the constant phase difference. Figure 1 illustrates this principle.

Allpass networks are distinctive since their attenuation is constant but phase shift changes with frequency. Mathematically, the transfer function looks like:

$$T(s) = \frac{(s - z_1)(s - z_2)(s - z_3) \ldots (s - z_n)}{(s + p_1)(s + p_2)(s + p_3) \ldots (s + p_n)}$$

In order to achieve the constant attenuation, the poles and zeros must be symmetric with the $j\omega$ axis. In other words, for any pole-zero pair:

the zeros are $\sigma + j\omega$ and $\sigma - j\omega$

and the poles are $-\sigma + j\omega$ and $-\sigma - j\omega$

The trick to achieve the constant phase difference is through proper placement of pole and zero locations. It's simple to say, "by proper placement of pole and zero locations...," but how is it really done?

Reference 2 goes to great lengths to develop the mathematical details, but only a synopsis is presented here. Consider two allpass networks which have phase shift functions (tan $(\theta_1/2)$ and tan $(\theta_2/2)$), then there is a function, tan $(\theta/2)$ which represents the phase difference where θ is $\theta_1 - \theta_2$. If the difference function is equated to a Cauer-Chebyshev polynominal, the roots of the polynominal are then the roots of the network transfer function. It so happens that roots of the Cauer-Chebyshev polynominal are elliptic functions which are solved in the design procedure.

Fortunately, with the aid of spreadsheets and up-to-date circuit implementations, it is possible to design realizable networks in minimum time. The design procedure for synthesis of networks producing a constant phase difference over a range of frequencies is shown below. Figure 3, a Lotus 123™ spreadsheet, automates the design.

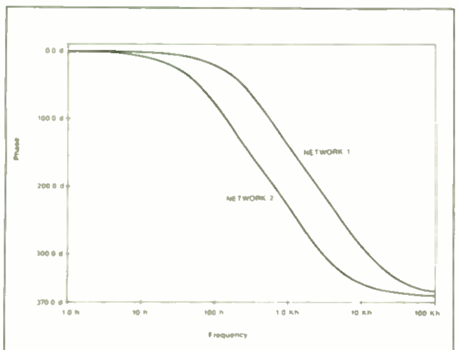

Figure 1. Allpass network phase transfer functions.

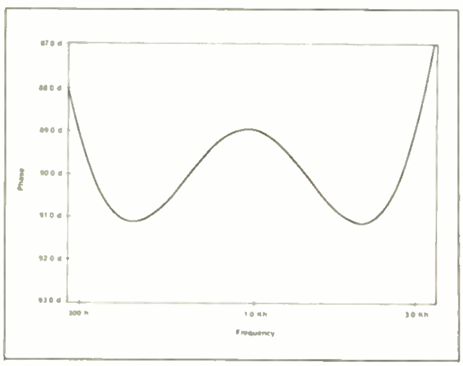

Figure 2. Phase difference plot: N =4, frequency ratio =10.

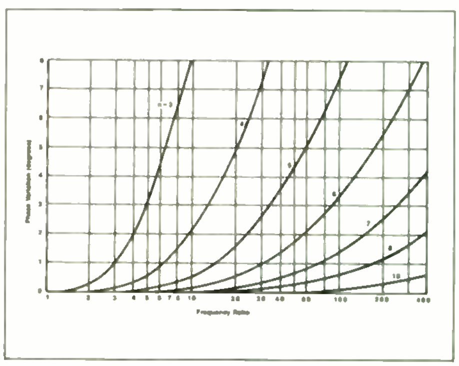

Figure 4. Phase variation for n-section networks.

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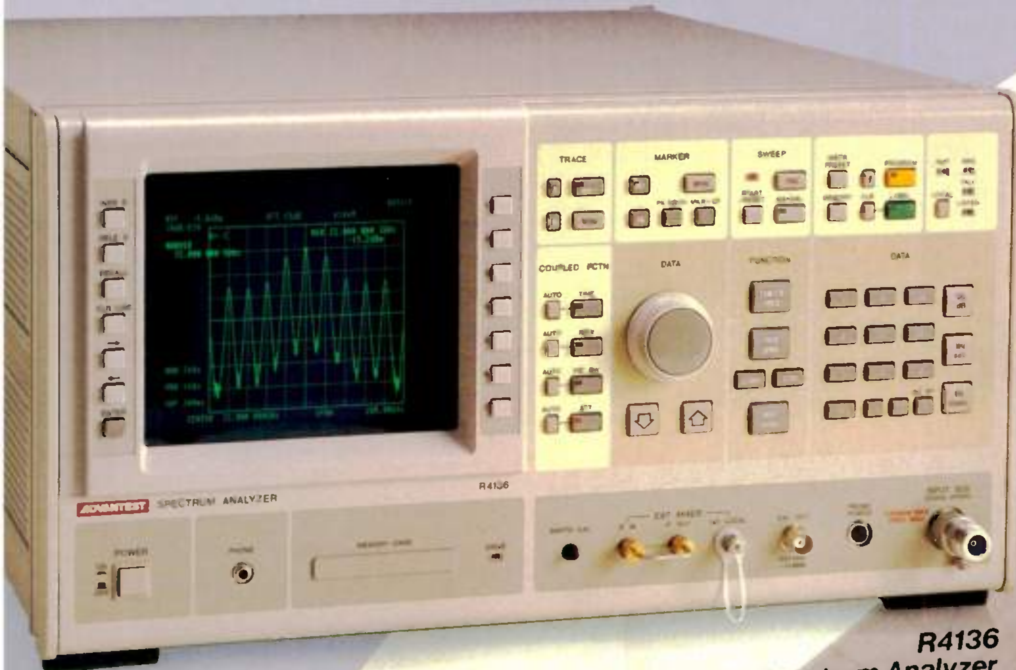

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```
A1: 'PHASE SHIFTER DESIGN SPREADSHEET
A3: 'ENTER HIGH FREQ LIMIT (HZ)
D3: [W12] 3000
A5: 'ENTER LOW FREQ LIMIT (HZ)
D5: [W12] 300
A7: 'ENTER NUMBER OF P-Z PAIRS
D7: [W12] 4
A9: 'FREQ RANGE IS
D9: [W12] +D3/+D5
A11: 'k IS
D11: (F4) [W12] @SQRT((D3^2)-(D5^2))/D3
E11: '
A13: 'k'IS
D13: (F4) [W12] @SQRT(1-D11^2)
A15: '==============================================
A17: 'APPROXIMATE q
A19: 'q0 IS
D19: [W12] 0.5*((1-@SQRT(D13))/(1+@SQRT(D13)))
A22: 'q IS
D22: [W12] +D19+(2*D19^5)+(15*D19^9)+(150*D19^13)
A24: '==============================================
A26: 'CALCULATE RHO POSTION
D26: [W12] '(SET UP FOR N=6)
A28: 'RHO 0
D28: [W12] (-@PI/4)/D7
A30: 'RHO 1
D30: [W12] (0.75*@PI)/D7
A31: '
A32: 'RHO 2
D32: [W12] (1.75*@PI)/+D7
A33: '
A34: 'RHO 3
D34: [W12] (2.75*@PI)/+D7
A35: '
A36: 'RHO 4
D36: [W12] (3.75*@PI)/+D7
A37: '
A38: 'RHO 5
D38: [W12] (4.75*@PI)/+D7
A40: '==============================================
A42: 'CALCULATE ROOTS
A44: 'MEAN FREQ (HERTZ)
D44: (S4) [W12] @SQRT($D$3*$D$5)
A45: 'MEAN FREQ (RADIANS)
D45: (S4) [W12] @PI*2*@SQRT($D$3*$D$5)
E45: '
D46: (F1) [W12] '
A47: 'p0
D47: (S3) [W12]
+D45*(@COS(D28)+((D22^2)*@COS(3*D28))+((D22^6)*@COS(5*D28)))/(@SI
N(D28)-((D22^2)*@SIN(3*D28))+((D22^6)*@SIN(5*D28)))
A49: 'p1
D49: (S3) [W12]
+D45*(@COS(D30)+((D22^2)*@COS(3*D30))+((D22^6)*@COS(5*D30)))/(@SI
N(D30)-((D22^2)*@SIN(3*D30))+((D22^6)*@SIN(5*D30)))
A51: 'p2
D51: (S3) [W12]
+D45*(@COS(D32)+((D22^2)*@COS(3*D32))+((D22^6)*@COS(5*D32)))/(@SI
N(D32)-((D22^2)*@SIN(3*D32))+((D22^6)*@SIN(5*D32)))
A53: 'p3
D53: (S3) [W12]
+$D$45*(@COS(D34)+(($D$22^2)*@COS(3*D34))+(($D$22^6)*@COS(5*D34))
)/(@SIN(D34)-(($D$22^2)*@SIN(3*D34))+(($D$22^6)*@SIN(5*D34)))
```

Figure 3. Lotus spreadsheet for allpass networks.

The procedure is:

1. Determine the desired phase shift (θ). In most applications this is $\pi/2$ or 90 degrees.

2. Determine the range of frequencies over which the constant phase shift is required. Let ω_1 represent the upper limit and ω_2 represent the lower limit. Compute the Frequency Ratio = ω_1/ω_2.

3. Determine the design criteria for an acceptable error bound for the Phase Variation (δ). The phase variation is the delta from the target phase shift value.

4. Using the chart in Figure 4, determine the number of sections (n) required to meet the Frequency Ratio and Phase Variation specifications.

5. Now compute the roots of the elliptic functions using the Fourier series expansions. Let Pσ represent the roots.

$$P_\sigma = (\sqrt{\omega_1\omega_2}) \frac{(\cos(\Gamma_\sigma) + q^2\cos(3\Gamma_\sigma) + q^6\cos(5\Gamma_\sigma) \ldots)}{(\sin(\Gamma_\sigma) - q^2\cos(3\Gamma_\sigma) + q^6\cos(5\Gamma_\sigma) \ldots)}$$

where $\sigma = 0, 1 \ldots (n-1)$

In order to calculate q, a number of constants must be determined:

a. the modulus $k = \sqrt{\omega_1^2 - \omega_2^2}/\omega_2$

b. the complementary modulus $k' = \sqrt{1 - k^2}$

c. The modular constant $q = e(-\pi K'/K)$. The modular constant q can be found by evaluating k'and k numerically, but there is a simpler approximation.

$$q \approx q_o + 2q_o^5 + 15q_o^9 + 150q_o^{13}$$

where $q_o = (1/2)(1 - \sqrt{k'})/(1 + \sqrt{k'})$

The angles, Γ_σ, can be calculated from:

$\Gamma_\sigma = (\sigma\pi - (1/2)\theta)/n$ radians

where $\sigma = 0, 1, \ldots (n - 1)$

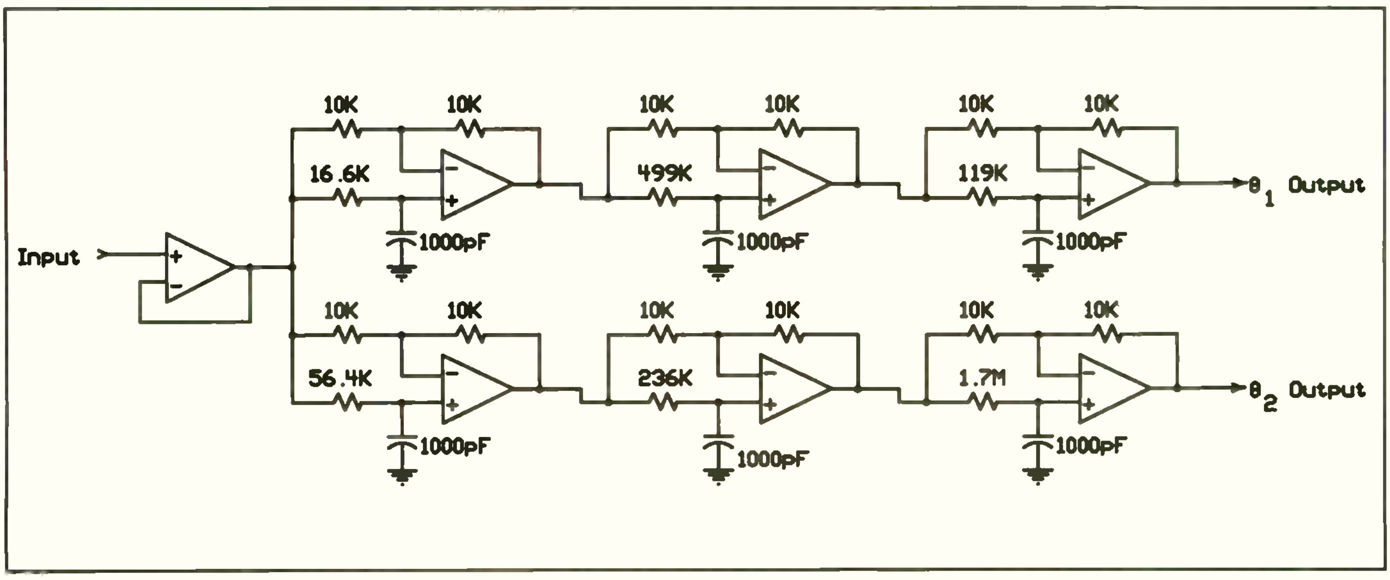


Figure 5. Example 1 network.

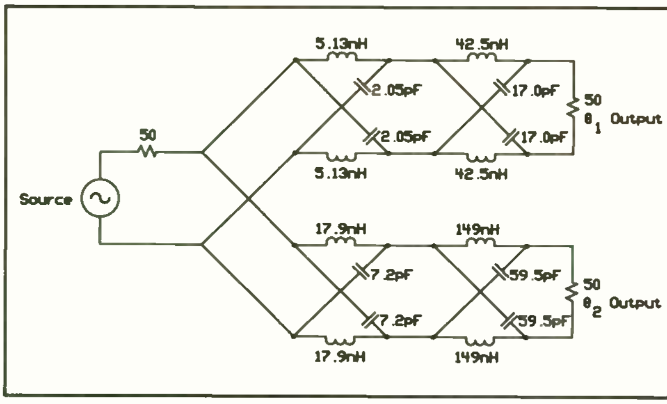


Figure 6. LC network from Example 2.

6. Group roots of like signs together. The negative roots will become one half of the network and the positive roots will become the other half. With the roots determined, a realizable circuit can be implemented to provide constant phase shift over the specified bandwidth.

Design Examples

Example 1. This circuit provides an audio phase shift network for use in a phasing type single sideband transceiver, with these design specifications:

Desired phase shift = 90°
Upper frequency limit = 3000 Hz
Lower frequency limit = 300 Hz
Phase Variation ≈ 1°

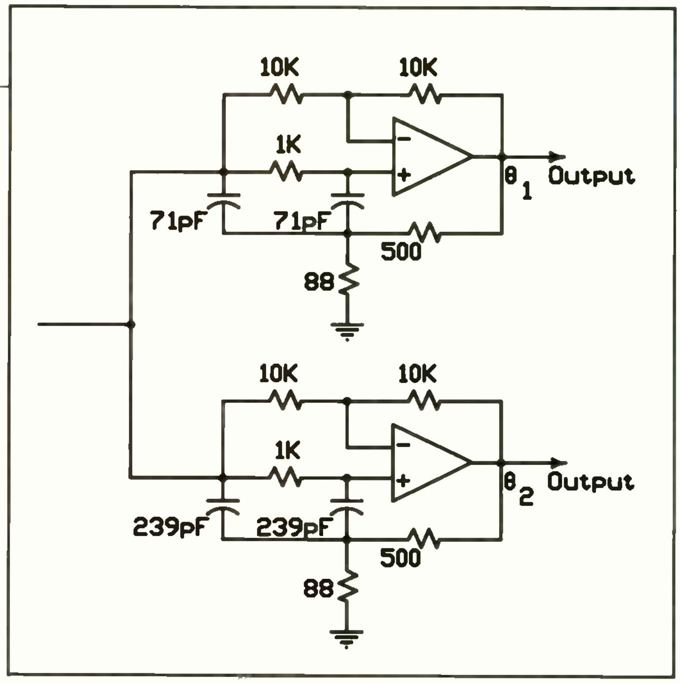
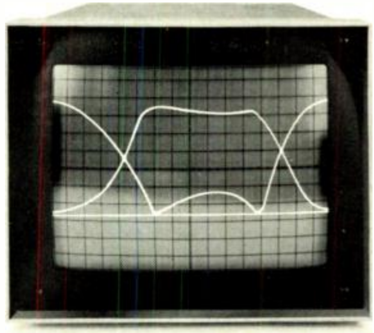


Figure 7. Steffan allpass network from Example 3.

By matching the Frequency Ratio (10) to the Phase Variation (1°), the closest "n" is equal to six. The upper branch of the circuit will have three pole-zero pairs and the lower branch will have three pole-zero pairs.



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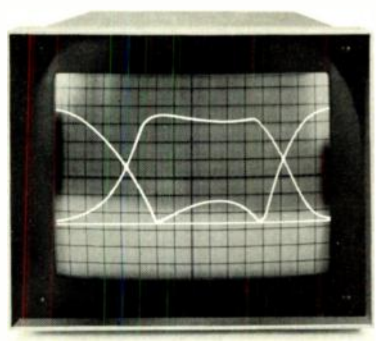

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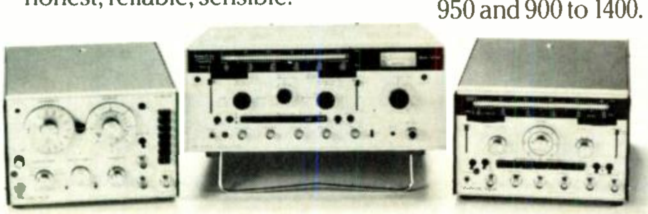

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From the spread sheet, the roots are:

p_o =-6.024E4
p_1 = 1.773E4
p_2 = 4.239E3
p_3 = 5.898E2
p_4 =-2.004E3
p_5 =-8.383E3

The negative roots will form the lower half of the circuit.

The circuit can be realized with the simple RC and op amp topology shown in Figure 5. Computation of the RC values is straight forward from the roots.

Example 2. Design an LC lattice network that meets the following specifications:

Desired phase shift = 90°
Upper frequency limit = 500 MHz
Lower frequency limit = 166 MHz
Phase Variation ≈ .2°

By matching the Frequency Ratio (3) to the Phase Variation (.2°), "n" equal to four is chosen.

From the spread sheet, the roots are:

p_o =-9.742E9
p_1 = 2.787E9
p_2 = 3.364E8
p_3 =-1.176E9

```
B1:  [W12] 'STEFFAN APN DESIGN SPREADSHEET
F1:  'TK
F2:  "4 NOV 88
A3:  [W12] "ENTER W1 =
B3:  (S3) [W12] 39370
B4:  (S3) [W12] '
A5:  [W12] "ENTER W2 =
B5:  (S3) [W12] 3559
A7:  [W12] "WO =
B7:  (S3) [W12] @SQRT(+B3*+B5)
A9:  [W12] "q =
B9:  (S3) [W12] +B7/(B3+B5)
A11: [W12] "ENTER R1 =
B11: (S3) [W12] 100000
C11: "R2 =
D11: (F0) +$B$25
A13: [W12] "b =
B13: (S3) [W12] -(4+B9^-2)/B11
C13: "R3 =
D13: +$B$23
A15: [W12] "c =
B15: (S3) [W12] 4/B11^2
C15: "C =
D15: (S2) (+B7^-1)*(@SQRT(B11^-1*(D11^-1+D13^-1))
A17: [W12] "a =
B17: (S3) [W12] 1
A19: [W12] "k1 =
B19: (S3) [W12] (-B13+(@SQRT(B13^2-4*B15)))/2
A21: [W12] "k2 =
B21: (S3) [W12] (-B13-(@SQRT(B13^2-4*B15)))/2
A23: [W12] "R3 =
B23: [W12] +B11/2
A25: [W12] "R2 =
B25: [W12] +B23/(B23*B19-1)
```

Figure 8. Spreadsheet for Steffan network design.

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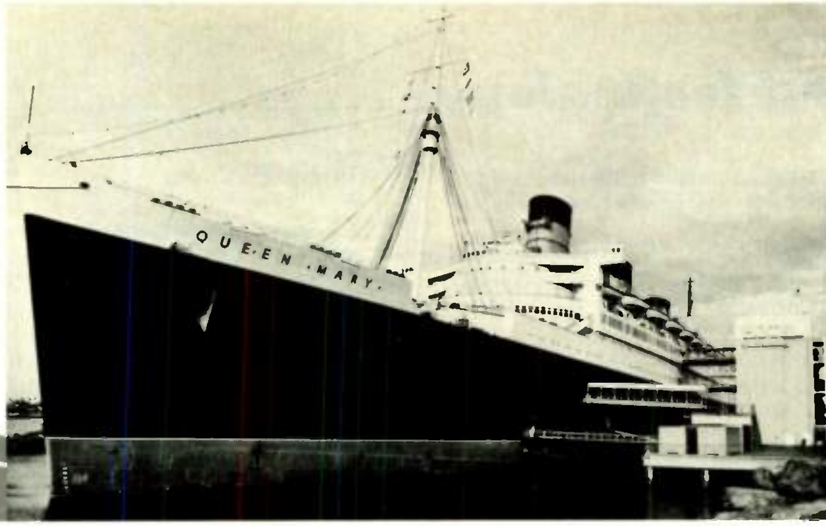

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Since the lattice configuration is specified, the resultant circuit is shown in Figure 6. Component values are computed from $L = R_o/p_o$ and $C = 1/(R_o p_o)$.

Example 3. The final example is for an RF phase shift network implemented using a high speed op amp such as the NE5539. Design specifications are:

Desired phase shift = 90°
Upper frequency limit = 5.0 MHz
Lower frequency limit = 4.0 MHz
Phase Variation ≈ .1°

This time, a Steffan allpass filter can be used for the upper and lower branches. Each Steffan filter has a second order transfer function, hence the complete transfer function of the network is fourth order and "n" also equals four.

From the spread sheet, the roots are:

p_0 =-1.417E8
p_1 = 4.210E7
p_2 = 5.572E6
p_3 =-1.875E7

The circuit equations to derive the component values for this filter are slightly more complex (see reference 6 for details). Figure 8 shows a spreadsheet automate the design process. The resultant circuit is shown in Figure 7.

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About the Author

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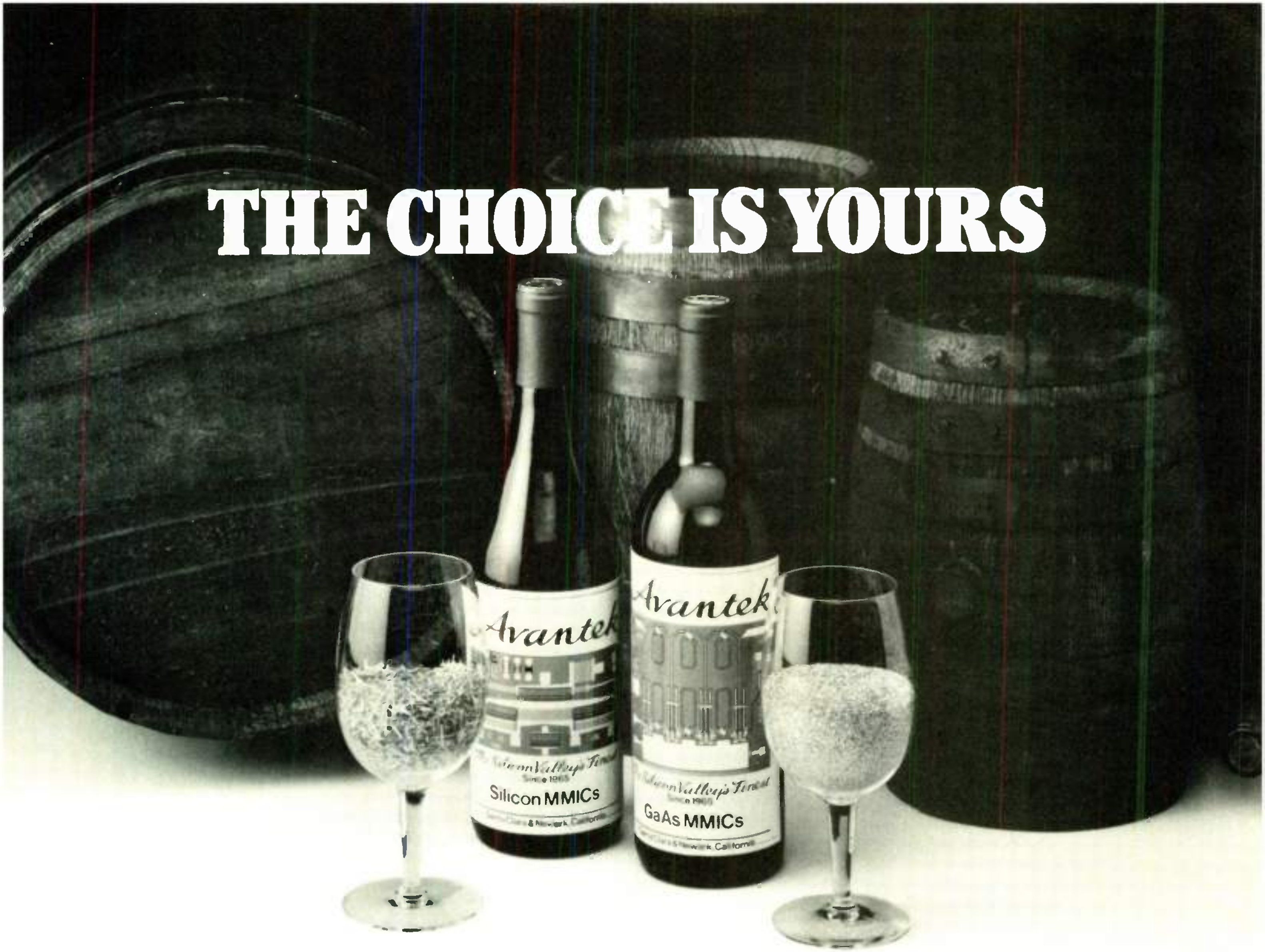

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MSA-0910	2.0	8.0	6.0	11.0
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MGA-61000	18.0	6.0	6.5	14.0

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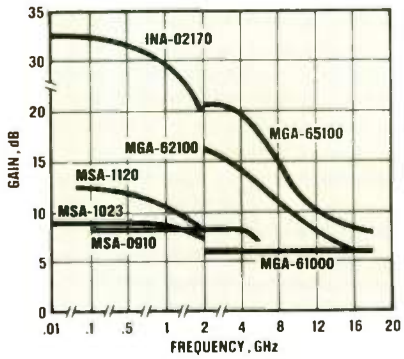

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BASIC Program for 90-Degree Allpass Networks

Quickly calculates up to 50 Phase-Shift Network Pole Pairs

By Allan G. Lloyd
Magnavox Electro-Optical Systems

This article describes a BASIC program developed to calculate the poles for 90-degree allpass networks. Like the previous article by Thomas Keely, the calculations are based on elliptic functions, but with some variation in calculation method.

A minimum specification for calculating the zero/pole frequencies for a 90 degree phase difference network could be:

F1 - The lower frequency limit, Hz.
F2 - The upper frequency limit, Hz.
N - The number of zero/pole pairs.

where: F2 > F1 > 0, and N = integer (1, 2, 3, etc).

From these inputs, range and center frequency are calculated, R = F2/F1 and FC = SQR(F1*F2). In the program described here, given any two of F1, F2, R or FC as input, the remaining two are calculated automatically. Note that range R is a dimensionless ratio, not a frequency difference, and FC is the geometric mean. R and N determine the remaining two parameters:

TOL = Phase tolerance, ± DEG
DB = Sideband suppression when used in an SSB modulator.

An early version of the program (1) was written for the now obsolete HP 2114 mini-computer. The program has since been translated into APPLE II Applesoft, Commodore BASIC, Datapoint BASIC, HP 1000 BASIC, IBM BASIC, and Borland Turbo BASIC. The version shown is in GWBASIC, which makes program segments easily referenced and explained by line number.

Operation

The program operates in an infinite loop which is exited by typing "X" at any prompt (line 2020). The input and output screens are shown in Figure 1. The input variables have a calculation priority based on their order of appearance on

```
                    90 DEG PHASE DIFFERENCE PROGRAM
                          By Allan G. Lloyd

            UP = up        <SP> = clear entry
            DN = dn        <CR> = enter value        X = exit program

input F1 =  300

input F2 =  3000

input FC =  948.6833

input R  =  10

input N  =  8

input TOL=  5.119102E-03

input DB = -86.99918

(Q=   .2621962 )

CALCULATE POLES NOW? - (Y/N/UP) _
```

(a) Input screen

```
A NETWORK, HERTZ          B NETWORK, HERTZ
------------------------------------------------
 12872.33                 3997.365
 2090.389                 1224.229
 735.1565                 430.5416
 225.1481                 69.91728

F1 = 300   HZ                           N   = 8   POLES
F2 = 3000   HZ                          TOL= +/- 5.136302E-03   DEG
FC = 948.6833   HZ                      DB = -86.97004
R  = 10
```

(b) Output screen

Figure 1. Program input/output formats.

the input screen, top to bottom. F1, if present, has highest priority, and DB, the lowest. Previously used values, if any, are stored and edited at the prompts. This saves having to retype them if only one input is changed for a new calculation. To change an input, type in a new value. To suppress an input, which forces recalculation of that input from lower priority inputs, press the space bar as input, or input a "0" (zero). Use the up/down arrow keys to move the cursor up/down to the next input position. This is very handy for exploring "what if" scenarios, or for correcting input errors.

For example, to design a speech bandwidth network for an SSB modulator, type "300" <CR> at F1 and then "3000" <CR> at F2. The program automatically calculates and prints FC = 948.6833, R = 10 and places the cursor at input N. Now type "8" for input N. Following this, TOL = 5.119102E−03 and DB = -86.99918 are calculated and printed. A prompt line appears asking if you want to calculate the poles now. If "N" is entered, the cursor restarts at F1, allowing any changes to be entered. If "Y" is entered, the pole frequencies are calculated, and the output screen appears. It prints the A-network and B-network frequencies in hertz, and displays the input parameters. Another prompt line appears, asking if you want this output to go to the printer, programmed as LPT1:.

The program will handle all 2^7 = 128 possible combinations of input. For example, input "space" for F1, F2, FC and R. Then input "4" for N and "1" for TOL. This is enough to calculate a normalized design whose center frequency is 1 (unit). (The program outputs F1=F2=FC=0, however.) Don't forget that the program uses variables in the order that they appear on the list. To use DB as an input after F1 and F2 are input, then both N and TOL must be blanked out (input as spaces or 0). The blanks are then reverse calculated and filled in. The whole procedure is easier done than said. Input can be in scientific form, i.e., R = 1E10. The program allows for N from 1 to 50 and R up to 9.99999E10.

All but the first character of the input line can be edited using the backspace key. If you make a mistake here, press ENTER and start over. The input process extends from lines 40 to 1080. The elliptic function "nome", Q, is calculated in 390-560. One of the equations involved is (510):

$$L = \frac{1}{2} \bullet \frac{1 - (1 - K1 * K1)\,\hat{}\,(1/4)}{1 + (1 - K1 * K1)\,\hat{}\,(1/4)}$$

K1=1/R

For large range, K1=1/R is small. K4=K1*K1 is even smaller, so that (1-K4)^(1/4) is near unity. The numerator term can result in large computational errors because you are subtracting nearly equal quantities. To avoid this, the expression for L was expanded using binomial series expansion into its equivalent (infinite) series form. The first twelve terms are used to calculate L for K1=0.7 (420-480). The dividing line at K1=0.7 was determined experimentally by comparing the results of both calculations for various K1. At K1=0.7 they are equal to nine decimal places.

The program contains another unique calculation, namely, the inverse calculation of R from Q (760-860). This is required when range data is not available as input. Line 760 calculates Q from N and TOL. Lines 790-850 use the first twenty terms of the inverse series expansion of Q, which is itself expressed as a rapid converging truncated series, (520-540). See Reference 4 for details.

After satisfactory iteration of performance parameters on the first screen, the main pole calculating routine (1090) is entered with a "Y" answer at the prompt (1050). Here also, the first twenty terms of the series expansion form for the pole frequencies are used (1150-1200). 1210 gives the normalized poles P(J), which are then frequency scaled in 1240. Line 1220 collects the phase contribution of each pole at F1 as it is calculated (inside the J-loop) and adds up the resultant phase error. 1260 and 1270 calculate the actual tolerance and DB based on the above. Output routine 1280-1580 prints either to the screen (SCRN:) or the printer (LPT1:).

Double Precision

Line 5 specifies double precision calculations for all variables except integer loop counters J and K%. The result is a much closer agreement between GWBASIC and double precision TURBO BASIC for large R and N (R=1E9 and N=24). Double precision GWBASIC is invoked as "GWBASIC/D". Slow down was not noticed on a 12 MHz AT clone with an 80287 co-processor. For small R and N (i.e., 10 and 4) omit line 5 and start in the single precision default mode as "GWBASIC".

Final Comments

This program is a major improvement in terms of user friendliness and speed over the original (1). An attempt was made to reduce the large number of GOTO's in it, in the name of "good" programming practice, but the effort was deemed not worth it. The original prototype version had only F1, F2 and N as input. The inverse mode of calculation and the other bells and whistles were added later.

This program is available on disk from the RF Design Software Service (details on page 78). The GWBASIC version and a compiled version are included. (Program listing appears on p. 53.) rf

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About the Author

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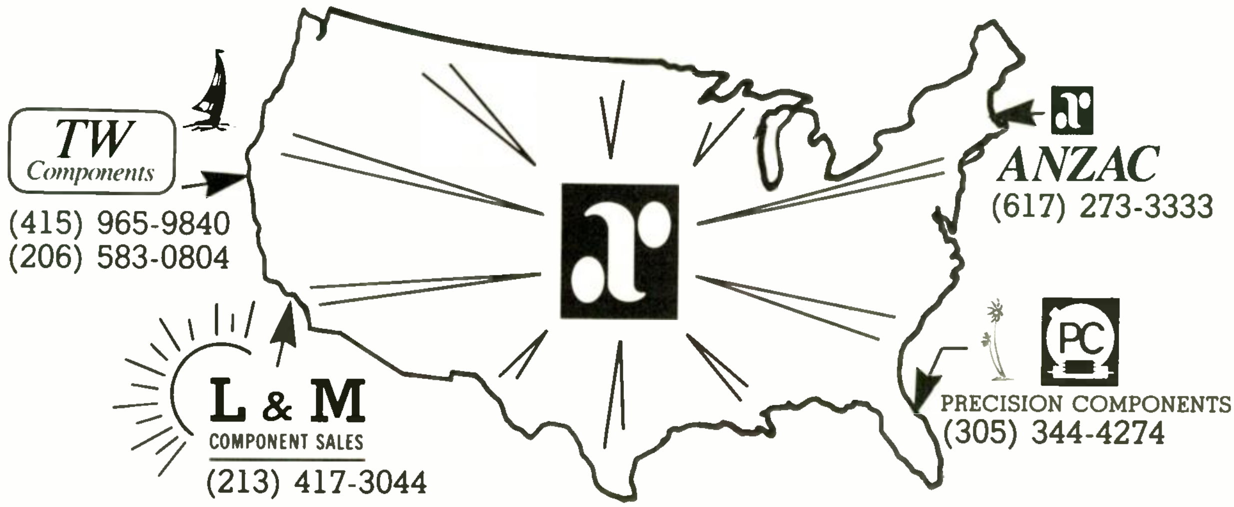
RF & MICROWAVE NEWS

April 7, 1989

ANZAC Announces Coast to Coast Distributors



Burlington, Ma. The *ANZAC* division of Adams-Russell recently announced the opening of three distributors in the United States. The introduction of one distributorship in Florida, two on the West Coast, and *ANZAC*'s own standard product distribution center in Massachusetts now makes local procurement of *ANZAC* catalog components easier than ever.

A spokesperson for *ANZAC* expounded on the advantages design engineers and procurement agents are receiving by dealing with their local *ANZAC* distributor. "By opening regional distributors, we now offer 2 distinct advantages to our customers. The first is local delivery. Each distributor is fully stocked with *ANZAC* components and can provide off-the-shelf delivery in 24 hours or less. The second advantage is service. *ANZAC* distributors have years of technical experience in the RF & Microwave industry and are already familiar with the *ANZAC* product line. They can offer technical assistance to design problems and provide the devices to solve those problems right away."

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```
1 ' AL90GW.TXT   -   90DEG PHASE DIFFERENCE NETWORKS  -
(GWBASIC)
2 '                          BY ALLAN G. LLOYD
3 '
4   DEFINT J : COLOR 7,0
10  DIM Z(50),P(50),W(50)
20  PI = 4*ATN(1): P2 = PI * PI: P7 = 720 / PI: LG = 20 /
LOG (10)
30  ON ERROR GOTO 2100

35  ' -----------------------------------------input F1
40  GOSUB 1800:GOSUB 2050 ' print heading & old values
50  FLAG = 0:MSG$ = "":VT = 7: GOSUB 1590  'recursion flag
60  ON G GOTO 50,80     'up or dn
70  F1 = PT             'assign input to variable

75  ' -----------------------------------------input F2
80  VT = 9: GOSUB 1590   'VT is vert tab, get input
90  ON G GOTO 50,160
100 F2 = PT
110  IF F1=0 OR F2=0  THEN 160   'input testing starts here
120  IF F1 < F2 THEN FC =  SQR (F1 * F2): GOTO 140
130 MSG$ = " REENTER F2": GOSUB 1690: GOTO 80
140 R = F2 / F1: IF R < 1E+11 THEN 380  'somewhat arbitrary
limit for R
150 MSG$ = " RANGE TOO BIG": GOSUB 1680: GOTO 50

155 ' -----------------------------------------input FC
160 VT = 11:G1 = G: GOSUB 1590      'save G for later
170 ON G GOTO 80,280
180 FC = PT
190  IF FC = 0 THEN 280
200  IF R >0 AND G1 = 1 THEN 340   'don't reenter R
210  IF F1 = 0 THEN 240
220  IF F1 < FC THEN F2 = FC * FC / F1: GOTO 140
230 MSG$ = " REENTER FC": GOSUB 1700: GOTO 160
240  IF F2 = 0 THEN 280
250  IF F2 > FC THEN F1 = FC * FC / F2:GOTO 140
260 GOTO 230
270 FLAG = 0

275  ' -----------------------------------input R (RANGE)
280 VT = 13: GOSUB 1590
290  ON G GOTO 160,570
300 R = PT
310  IF R = 0 THEN 560
320  IF (R > 1 AND R < 1E+11) THEN 340
330 MSG$ = " REENTER RANGE": GOSUB 1710: GOTO 280
340  IF FC >0 THEN F1 = FC /  SQR (R):F2 = FC *  SQR (R):
GOTO 380
350  IF F1 > 0 THEN FC = F1 *  SQR (R):F2 = F1 * R: GOTO 380
360  IF F2 >0 THEN F1 = F2 / R:FC = F2 / SQR (R): GOTO 380
370  GOTO 390
380  GOSUB 1680: GOSUB 1690: GOSUB 1700: GOSUB 1710  'print
F1, F2, Fc, R

385  ' --------------------------------------CALCULATE 'Q'
390 K1 = 1 / R
400 K4 = K1 ^ 2
410  IF K1 > .7 THEN 500   'dividing line for direct or
series calc of Q
420 Y = 0                  'where results are equal to 9
digits
430  FOR J = 12 TO 1 STEP - 1    'series calc
440 Y = (J - 1.25) * K4 * (1 + Y) / J
450  NEXT J
460 L = - Y / (4 + 2 * Y)
470  IF L > .000001 THEN 520
480 Q1 = L
490  GOTO 540
500 K2 = (1 - K4) ^ (1 / 4)       'direct (closed form) calc
510 L = .5 * (1 - K2) / (1 + K2)
520 M = L ^ 4
530 Q1 = L * (1 + M * (2 + 15 * M))
540 Q = EXP (P2 / LOG (Q1))        'a really weird formula!
550  LOCATE 21,1: PRINT "(Q= ";Q;")"
560  IF FLAG =1 THEN GOSUB 1720: GOSUB 1730: GOSUB 1740:
GOTO 1030: 'OUTPUT

565  ' ------------------------------------------input N
570 VT = 15: GOSUB 1590
580  ON G GOTO 270,700
590 N = PT
600 N% = INT ( ABS (N) + .5)  'round up
605 LOCATE 15,30: GOSUB 2040  'blank old N1
610  GOSUB 1720  'print N
620  IF N% < 51 THEN 640
630 MSG$ = " TOO BIG, RE-ENTER": GOSUB 1720: GOTO 570
640  IF N%=0 OR R=0 THEN 700
650 T = P7 * Q ^ N%                 'calc prelim TOL from Q & N
660 DB = LG * LOG ( TAN (2 * Q ^ N%))    'calc DB from Q & N
670  GOSUB 1730: GOSUB 1740    'PRINT TOL & DB
690  GOTO 1030  'go test for pole calc

695  ' ------------------------------------input T  (TOL)
700 VT = 17: GOSUB 1590
710  ON G GOTO 570,950
720 T = PT
730  IF T = 0 THEN 950
740 DB = LG *  LOG ( TAN (PI * T / 360)): GOSUB 1740
'calc DB from T
750  IF N% = 0 THEN 890
760 Q = (T / P7) ^ (1 / N%)            'calc Q from N
770 Y3 = .5: Y4 = .5
780 C = 0
790  FOR J = 1 TO 20                   'calc R from Q
(inverse calc)
800 C = C + 2 * J - 1
810 E = Q ^ C
820 Y3 = Y3 + E
830 Y4 = Y4 + E * ( - 1) ^ J
840  IF E < .0000001 * Y4 THEN 860  'prevented underflow in
old HP comp.
850  NEXT J
860 R = (Y3 / Y4) ^ 2
870  GOSUB 1710:  'print R   'recursion here. after getting
R, go back
880 FLAG = 1: GOTO 310        'and calc F1, F2, & Fc (if
possible)
890  IF R = 0 THEN 40
900 N1 = LOG (Q1) * LOG (T / P7) / P2    'calc theory N from
Q & TOL
910 N% = INT (N1 + .999999)              'round up
920  GOSUB 1720:  'PRINT N%
930  LOCATE 15,30: PRINT "(N1= ";N1;")"   '(for info only)
940  GOTO 1030

945  ' -----------------------------------------input DB
950 VT = 19: GOSUB 1590
960  ON G GOTO 700,50
970 DB = - ABS (PT)     'always expressed as neg
980  GOSUB 1740
990  IF DB = 0 THEN 50    'restart
1000  T = (360 / PI) * ATN ( EXP (DB / LG))  'calc T from DB
1010  GOSUB 1730    'print TOL
1020 GOTO 750       'go back for N

1030  LOCATE 23,40: GOSUB 2040  'blank old msg (from next
line)
1035  IF FC = 0 THEN  PRINT "(NORMALIZED ONLY)";
1040 LOCATE 23,1: PRINT "CALCULATE POLES NOW? - (Y/N/UP) ";:
GOSUB 2000

1050  IF X$ = "Y" OR X$ = "y" THEN 1090
1060  IF X = 72 THEN 950   'UP
1070  IF X = 80 THEN 50    'DN
1080  GOTO 50              'default restart

1085 '-----------------this is the pole calculating routine
1090 S = PI / 2
1100  PRINT : PRINT "CALCULATING POLE #";
1110  FOR J = 1 TO N%
1120  PRINT " "; J;
1130 U = 0: V = 0
1140 Z(J) = ( - 1) ^ (J - 1) * PI * (2 * J - 1) / (4 * N%)
1150  FOR K% = 0 TO 20                     'uses series
approximation
1160 W = (Q ^ (K% + 1)) ^ K%
1170  IF W < 9.999999E-21 THEN 1210
1180 U = U + W * COS ((2 * K% + 1) * Z(J))
1190 V = V + ( - 1) ^ K% * W * SIN ((2 * K% + 1) * Z(J))
1200  NEXT K%
1210 P(J) = ( - 1) ^ N% * U / V            'normalized poles
1220 S = S + 2 * ATN ( SQR (K1) / P(J))   'add up phase
error
1230 P(J) =  ABS (P(J))                   'get rid of sign
1240 W(J) = FC * P(J)                     'scaled poles
1250  NEXT J: PRINT : PRINT

1260 T = 180 * S / PI                     'actual TOL based
on sum at F1
1270 DB = LG * LOG ( TAN (S / 2))         'for SSB

1275 '-----------------------------------output routines
1276 '  "#1" = SCRN: or (if selected) LPT1:
1277 '
1280  CLS: OPEN "O",1,"SCRN:"             'default to screen
1290  IF FC = 0 THEN 1370                 'print normalized
1300  PRINT #1, "A NETWORK, HERTZ"; TAB(26); "B NETWORK,
HERTZ"
1310  PRINT #1,
"---------------------------------------"
1330  FOR J = 1 TO N% STEP 2
1340  IF J = N% THEN  PRINT #1, W(J): GOTO 1440    'print
last odd pole
1350  PRINT #1, W(J); TAB(25); W(J + 1)
'alternate poles to A & B
1360  NEXT J: GOTO 1440

1370  PRINT #1, "A (normalized)"; TAB(26): "B (normalized)"
1380  PRINT #1,
"---------------------------------------"
1400  FOR J = 1 TO N% STEP 2
1410  IF J = N% THEN  PRINT #1, P(J): GOTO 1440
1420  PRINT #1, P(J); TAB(25); P(J + 1)
1430  NEXT J

1440  PRINT #1,   'print summary of inputs
1450  PRINT #1, "F1 ="; F1; " HZ"; TAB(35); "N  ="; N%; "
POLES"
1460  PRINT #1, "F2 ="; F2; " HZ"; TAB(35); "TOL= +/-"; T; "
DEG"
1470  PRINT #1, "FC ="; FC; " HZ"; TAB(35); "DB = "; DB
1480  PRINT #1, "R  ="; R
1520  PRINT #1,
1550  CLOSE  '(SCRN or LPT1)

1560  LOCATE CSRLIN,1:PRINT "WANT PRINTOUT - (Y/N)? ";:
GOSUB 2000
1570  IF X$="Y" OR X$="y" THEN OPEN "O",1,"LPT1:":GOTO 1290
'print it!
1580  GOTO 40  'otherwise, restart using old values

1585  ' --------------------------------input subroutine
1590  LOCATE VT,12: G=0: GOSUB 2000
1610  IF X = 72 THEN G = 1: RETURN :  'up=H
1620  IF X = 80 THEN G = 2: RETURN :  'down=P
1630  IF X = 32 THEN PT = 0: GOTO 1670: '<SPACE>
1640  IF X = 13 THEN ON (VT - 5) / 2 GOTO
1680,1690,1700,1710,1720,1730,1740
1650  PRINT X$;: INPUT "", A$          'print 1st char &
input the rest
1660 A$ = X$ + A$: PT = ABS(VAL (A$))  'combine & remove
sign
1670  GOTO 1750  'print it

1680 VT = 7:PT = F1: GOTO 1750    'print old values
1690 VT = 9:PT = F2: GOTO 1750
1700 VT = 11:PT = FC: GOTO 1750
1710 VT = 13:PT = R: GOTO 1750
1720 VT = 15:PT = N%: GOTO 1750
1730 VT = 17:PT = T: GOTO 1750
1740 VT = 19:PT = DB

1750  LOCATE VT,12: GOSUB 2040    'blank old value
1770  IF MSG$ <> "" THEN PRINT MSG$: MSG$ = "": RETURN
1780  IF PT = 0 THEN MSG$ = " ": GOTO 1770    'val=0 used in
branching tests
1790  PRINT PT: RETURN

1795  '  ---------------------------------- print heading
1800 CLS : PRINT  TAB(20);"90 DEG PHASE DIFFERENCE PROGRAM"
1805 PRINT TAB(27);"By Allan G. Lloyd"
1810  PRINT : PRINT  TAB(13);"UP = up      <SP> = clear
entry
1820  PRINT  TAB(13);"DN = dn      <CR> = enter value";
1825  PRINT SPC(8); "X = exit program"
1830  LOCATE 7,1
1835  PRINT "input F1 = ": PRINT
1840  PRINT "input F2 = ": PRINT
1850  PRINT "input FC = ": PRINT

1860  PRINT "input R  = ": PRINT
1870  PRINT "input N  = ": PRINT
1880  PRINT "input TOL= ": PRINT
1890  PRINT "input DB = ": PRINT
1900  RETURN

1910 ' get input char
2000  COLOR 23: PRINT "_";:COLOR 7   'blinking underline
2010  X$=INKEY$:IF LEN(X$)=0 THEN 2010   'test for keypress
2020  IF X$="X" OR X$="x" THEN CLOSE:CLS:PRINT"EXIT":END
'exit
2030   X=ASC(RIGHT$(X$,1)):LOCATE CSRLIN,POS(0)-1:PRINT "
";:RETURN

2035   'blanking subroutine
2040 FOR J=1 TO 30: PRINT " ";: NEXT: LOCATE
CSRLIN,POS(0)-30:RETURN

2045   'print variables
2050 GOSUB 1680:GOSUB 1690:GOSUB 1700:GOSUB 1710:GOSUB 1720
2060 GOSUB 1730:GOSUB 1740: RETURN  'print F1, F2, Fc, R, N,
TOL, DB

2100 RESUME 40 ' error handler
```

Figure 2. Program listing (GWBASIC).

Easy Phase-Noise Measurement

By Raymond Dewey
Sprague Electric Co., Semiconductor Group

The spectral purity of radio-frequency sources has become a significant concern in the design of radio communications and certain radar systems. In response to this relatively new requirement, phase-noise measurements have become an important method of verifying the purity and short-term stability of these sources. Presented here is an easily constructed phase-noise measurement circuit using an inexpensive, readily available consumer integrated circuit.

Originally designed as an application-specific integrated circuit to decode the C-QUAM[R] AM stereo signal (1), the MC13020D/P (Motorola or Toshiba) or ULN3820A/LW (Sprague) can also be used to measure the phase noise of a signal source with significant reductions in cost and complexity. (Motorola, Toshiba, and Sprague manufacture the same IC under a cross-licensing agreement.) A simplified functional block diagram of this IC is shown in Figure 1. Since the L-R component of the C-QUAM stereo signal is phase-modulated, the integrated circuit contains a phase-locked loop, an envelope detector, synchronous in phase (I) and quadrature (Q) phase detectors, as well as other circuitry to process and detect the 25 Hz phase-modulated pilot signal used in C-QUAM AM stereo broadcasting.

To get some idea of the requirements of this complex consumer IC AM stereo decoder, from Reference 2, consider that it has an audio signal-to-noise ratio of about 55 dB at 30 percent AM stereo modulation. The peak phase modulation of the carrier at 30 percent L-only modulation is 23.2 degrees, so the residual phase modulation of either the decoder PLL or incoming signal must be less than 0.073 degree RMS (0.106 degree peak). The 25 Hz level at the quadrature detector is typically 20 mVp, and the noise is about 1 mVp. This gives a typical phase noise of about 0.07 degree peak at the low frequencies of interest. Because the integrated circuit is designed to provide an output signal-to-noise ratio of greater than 50 dB in the stereo mode, the phase-locked loop is very quiet. The loop is normally designed for a natural frequency of about 10 Hz so that the 25 Hz C-QUAM AM stereo pilot can be recovered from the quadrature phase detector. Internal emitter-coupled logic dividers are used to insure good phase accuracy and low noise for the signals used in the phase detectors. The device is designed for an IF input of 450 or 455 kHz. The oscillator runs at eight times the input frequency, which ensures that it is outside the AM broadcast band, even if a 260 kHz IF frequency were used.

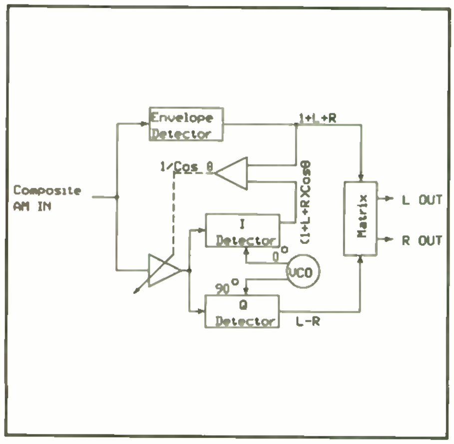

Figure 1. Simplified C-QUAM decoder function.

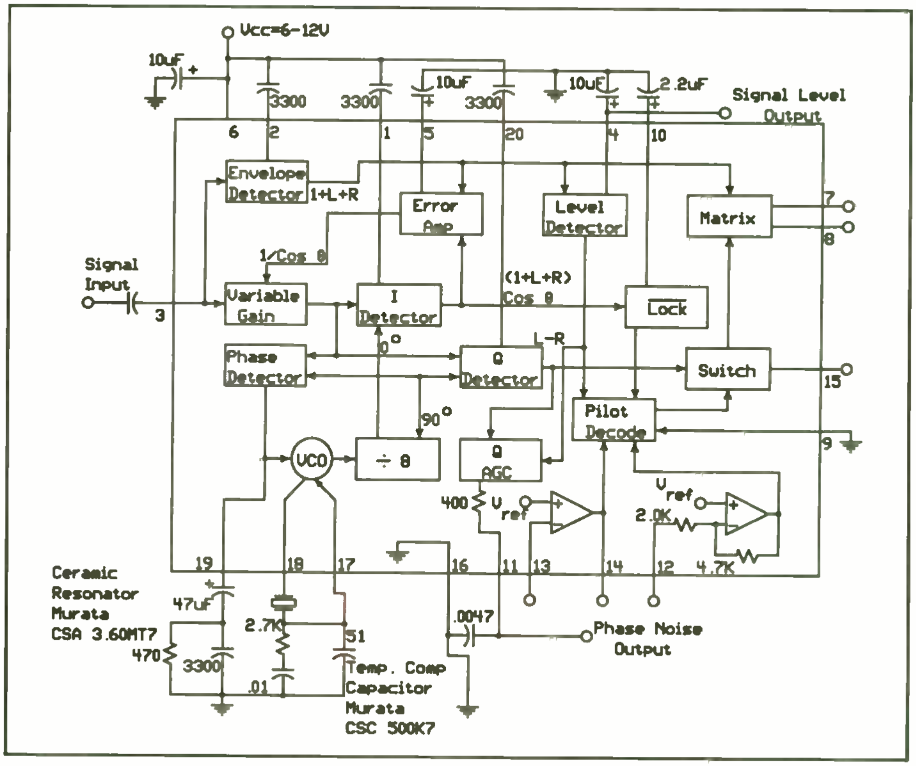

Figure 2. Phase noise detector.

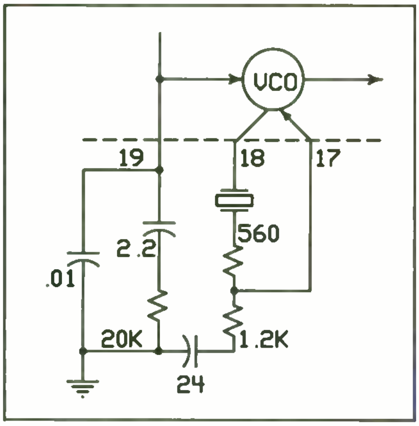

Figure 3. Crystal resonator.

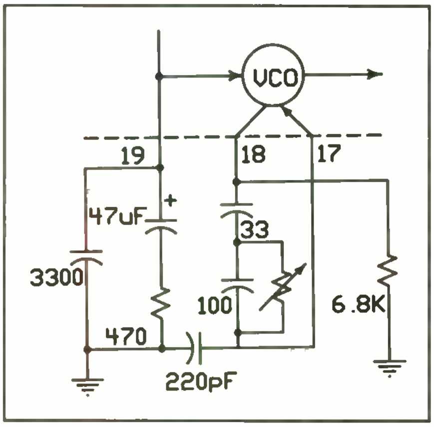

Figure 4. LC resonator.

Because the device has such a low loop cutoff frequency, the quadrature detector output (pin 11) can be used to observe the phase noise of the receiver local oscillator or any other input signal. A phase-noise detector using the MC13020D/P is shown in Figure 2. Alternative oscillator configurations using crystal or LC resonators are shown in Figures 3 and 4. The pull-in range for the crystal oscillator is about ± 100 Hz (at 450 kHz) and de-Q-ing with a series resistor can increase the range only slightly. For the LC and ceramic oscillators, the pull-in range is about ± 2500Hz (at 450 kHz). The VCO constant (K_o) is the conversion factor between the VCO frequency and control voltage in radians persecond. For the three oscillator types, the constants are:

Ceramic K_o = 30k to 40k (Figure 2)
Crystal K_o = 1250 (Figure 3)
LC K_o = 28k (Figure 4)

The crystal oscillator features the lowest inherent phase noise while the LC circuit has the advantage of being adjustable and can be used to center the circuit operation at the exact frequency. The signal input impedance at pin 3 is in excess of 20 K. A 200 mV RMS carrier level is an ideal input level, but because of the internal AGC circuit, the input level can be between 100 and 350 mV RMS. The source impedance at the level detector output (pin 4) is about 8.2 kohms and the output voltage will vary from about 1.6 VDC with no signal applied to a maximum of about 3.3 VDC.

With the components shown, the loop

C-QUAM AM Stereo

One of the simplest AM stereo methods would use quadrature modulation with two amplitude modulated RF carriers that are 90 degrees out of phase. L+R information would be on the in-phase carrier and L-R would be suppressed-carrier modulated on the quadrature carrier. However, about 14 percent distortion occurs in a standard AM envelope detector when this system is transmitting in stereo, making it incompatible with monophonic receivers.

The Motorola C-QUAM system provides compatibility by modulation both in-phase and quadrature carriers by the cosine of the modulation angle (cosϕ). The MC13020 IC uses full-wave envelope detection, a phase-locked loop for suppressed-carrier restoration, and a comparator and variable-gain amplifier to develop the cosine correction signal. The 1/cosϕ correction signal is obtained from the difference between the envelope detector and the I detector outputs, then applied to the variable-gain amplifier at the I and Q inputs. The Q detector and envelope detector outputs then feed the decoding matrix for stereo output.

The system also uses a 25 Hz pilot tone, added to the L-R signal at 4 percent modulation to identify stereo transmission. A signal level detector drives the Q AGC to provide a constant level to the pilot decoder. The same signal level information is supplied to the decoder. With a strong signal, the decoder switches to stereo in 7 cycles, but with a weaker signal, it takes 37 cycles before switching to avoid false triggering.

frequency is about 10 Hz so that the quadrature detector output will have a flat response from 25 Hz to about 7 kHz. The bandwidth will be limited by the bandwidth of any input tuned circuits and the value of the quadrature detector filter capacitor at pin 20. Of course, a lower loop natural frequency could be used if desired. It may be necessary to adjust the resistor in series with the loop filter capacitor at pin 19 to make the quadrature output flat down to the loop-cutoff frequency. The loop filter components can be calculated from:

$$C = K_d K_o/(2\pi f_n)^2$$

$$R = \zeta/\pi f_n C$$

Where,

K_d = Phase detector gain constant. The conversion factor between the phase detector output voltage and the phase difference between input and VCO signals in volts per radian.

ζ = Damping factor. The ability of the loop to respond quickly to an input frequency step without excessive overshoot. In this application it is recommended that ζ be kept between 0.5 and 0.8.

f_n = Natural frequency of the loop which is determined by the final pole positions in the complex plane. This is the low-frequency cutoff of the loop when the loop is critically damped (ζ = 0.707).

The phase detector output at pin 11 is 636 mVp per radian. The source impedance of pin 11 is approximately 400 ohms. Any phase noise produced by the signal source will show up here An operational amplifier between pins 13 and 14 can be used for other purposes. rf

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rfi/emc corner

A Fundamental Review of EMI Regulations

By Daryl Gerke
Kimmel Gerke Associates, Ltd.

With the increasing demand for electronic communications, the electromagnetic spectrum has become very crowded. Aggravating this problem is pollution from non-intentional users of electromagnetic energy, such as computers, microprocessors, and household appliances. To control this problem, various government regulations have emerged. This article compares some electromagnetic interference (EMI) regulations that apply to non-intentional users of radio frequencies. Although military regulations are discussed, the primary focus is on international commercial regulations.

Addressed in this article are five major sets of government regulations for EMI. These are MIL-STD-461/462 (United States, Department of Defense), MIL-E-6051 (United States, Department of Defense), FCC Part 15J (United States, Federal Communications Commission), VDE 0871 (West Germany, Verband Deutscher Electrotechniker), and VCCI (Japan, Voluntary Control Council for Interference). The first two sets of regulations apply to military equipment, while the remaining three apply to commercial equipment.

Any EMC situation can be described in terms of a source-path-victim model. Electronic equipment can be both a source of emissions and a victim susceptible to emissions. These emissions can be intentional, as with radio equipment, or unintentional, as with computers and microprocessors. The energy coupling paths can be either conducted, by wiring and components, or radiated, by electromagnetic fields.

These distinctions in coupling paths and directions have led to the classification scheme shown in Table 1, which accounts for both the status (source or victim) and the coupling path (conducted or radiated) of equipment. To assure electromagnetic compatibility, one must keep the radiated emissions (R_E) below the radiated susceptibility (R_S) thresholds, and the conducted emissions (C_E) below the conducted susceptibility (C_S) thresholds of the system.

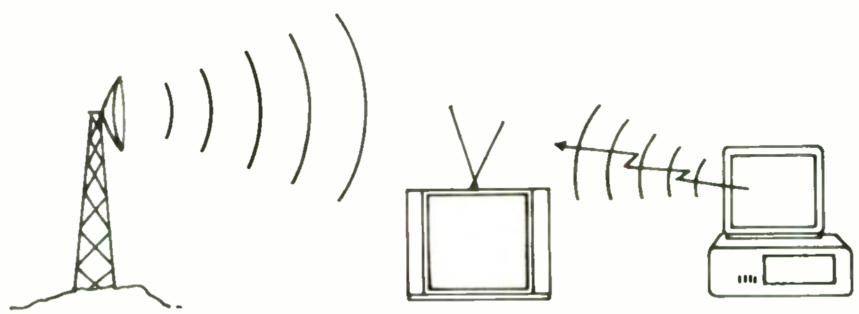
C_E	: Conducted emission
C_S	: Conducted susceptibility
R_E	: Radiated emissions
R_S	: Radiated susceptibility

Table 1. Classification scheme.

Military vs. Commercial Requirements

The differences in military and commercial EMI regulations are due both to the environments and to the system objectives. In general, military requirements are more stringent than commercial ones, and thus more difficult and more expensive to meet.

Military regulations are aimed at assuring mission success. The goal is to ensure satisfactory operation, or self-compatibility within a system, by maintaining margins between sources and victims. Thus, both susceptibility and emission levels are addressed, and the particular environment (ground, airborne, ship) is considered. Equipment processing classified data may be further constrained with the TEMPEST regulations. These regulations deal only with emissions that could compromise security.



E db μV/m (Expected useful signal) 300-1000 μV/m − S/N (Signal to Noise) 20-40 dB + Bldg atten 0-10 dB = Limit at distance (3 to 30 meters)

Figure 1. Regulatory model for radiated emissions.

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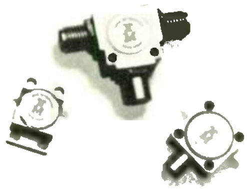

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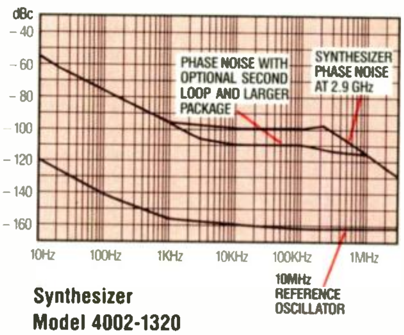

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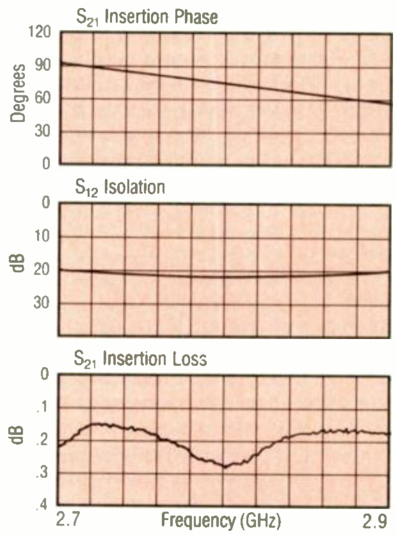

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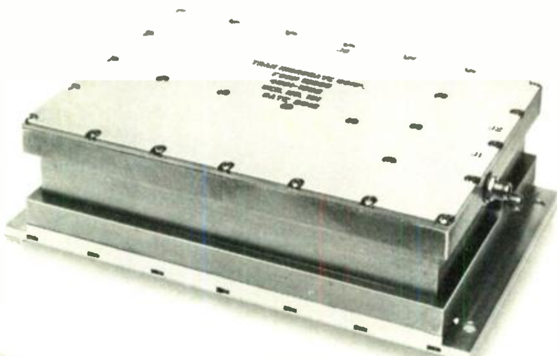

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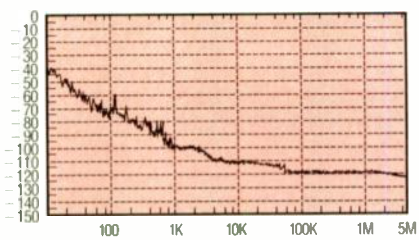

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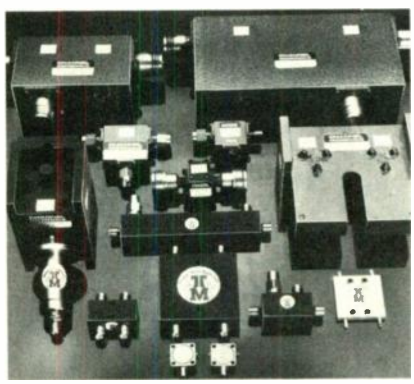

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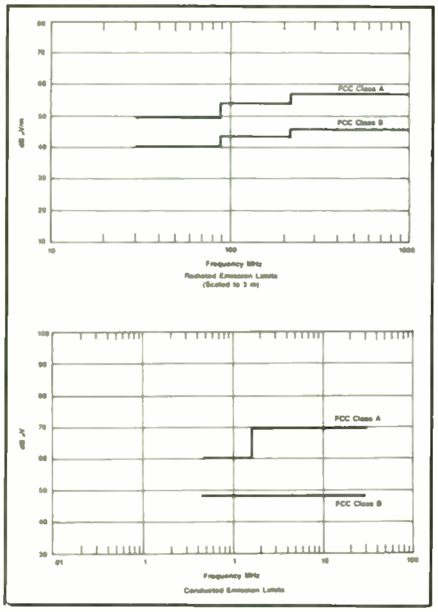

Figure 2. FCC limits for radiated and conducted emissions.

The military requirements include both equipment level and system level specifications. At the equipment level (MIL-STD-461/2), the levels and test methods are quite specific, while at the systems level (MIL-E-6051), the issue of overall compatibility is addressed. The underlying assumption is that if the individual units meet MIL-STD-461, the probability of system success is very high. In addition, specific tests and test levels can be waived or modified as part of the contract negotiations, assuming such modifications make sense.

Commercial regulations are aimed at protecting commercial communications equipment from pollution of the electromagnetic spectrum. The general intention is to limit emissions from equipment to levels that will not interfere with residential radio and television reception. Thus, only emissions are addressed. Furthermore, those levels are often divided into conducted emissions below 30 MHz and radiated emissions above 30 MHz, since conduction is more likely at low frequencies, and radiation is more likely at the higher frequencies.

The commercial requirements are not divided into equipment and system level categories. Rather, the underlying assumption here is that radio/television reception can be assured by keeping emission levels low enough. Incidentally, since the commercial radiated emissions regulations are essentially based on television signal levels, they may not be adequate to protect more sensitive land mobile radio communications receivers (5).

In the commercial world, tests and test levels cannot be waived or modified. In fact, in most instances, equipment cannot be sold until it is fully compliant with the appropriate regulations. Figure 1 shows a model for the commercial radiated emission requirements. The aim is to keep the emissions 20 dB or more below normal television signal levels in an urban environment. Assumptions on typical distances between computer and television have resulted in the 3, 10, and 30 meter test distances.

Although it is difficult to make a direct comparison between military and commercial specifications, military specifications typically require 30 to 60 dB more shielding to meet the limits that comparable commercial systems require.

MIL-STD-461

Due to its longevity and widespread use, this is a very influential EMC specification for individual electronic equipment, even in the commercial world. Originally issued in 1967, it has undergone numerous changes and refinements. The current version, MIL-STD-461C, was issued in 1986.

It is accompanied by two additional documents — MIL-STD-462, which defines the test methods, and MIL-STD-463, which defines units, abbreviations, and acronyms. MIL-STD-461 actually defines the test levels and limits.

Over the years, more than twenty different test procedures have been specified by these standards. Only some, however, will apply to any specific equipment. MIL-STD-461 groups equipment into general classes, depending on type and use (airborne, ground, ship, etc.), and then specific tests that apply are called out in tables. In addition, further modifications to specific test requirements may be part of contract negotiations.

MIL-E-6051

This specification describes overall requirements for EMC at the system, rather than at the equipment, level. As such, it includes control of the electromagnetic environment, lightning, static, bonding, and grounding. It also addresses electromagnetic hazards to ordnance and personnel. By its very nature, this specification is intended to apply to acquisition of major military systems. It is also often applied to subsystems, to assure coordination of EMC issues. It does reference MIL-STD-461 and MIL-

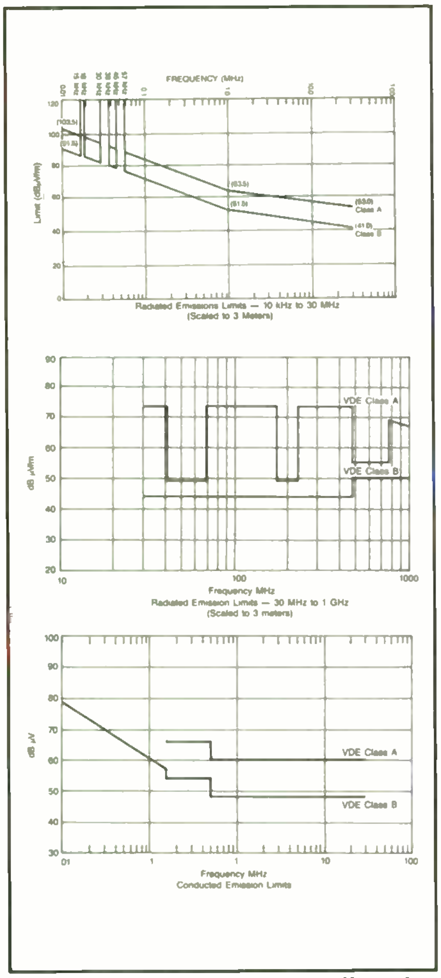

Figure 3. VDE limits for radiated and conducted emissions.

STD-462, as well as numerous other applicable specifications and standards, both military and commercial.

FCC Part 15J

In the United States, the Federal Communications Commission (FCC) is responsible for regulating devices which can cause interference to radio or television reception. This includes any industrial, commercial, consumer, or non-government user of the radio frequency spectrum in the United States.

The major parts of the FCC rules and regulations most applicable to EMC are set forth in Title 47, Parts 15, 18, and 68 of the U.S. Code of Federal Regulations. These are augmented by a series of bulletins that clarify various provisions of these regulations.

A major regulation of interest is Part 15, which sets the standards for unlicensed incidental and restricted radiation devices. Of particular interest is Subpart J of Part 15, which covers the EMI limits for computing devices. Due to the proliferation of computers in today's society, FCC Part 15J is probably the most visible and most notorious of today's EMC specifications.

Equipment subject to Part 15J is divided into two classes, based on type of use. Class A equipment is used solely in commercial environments, while Class B equipment can be used in either commercial or residential environments. The limits are emissions only, and are divided into conducted only below 30 MHz and radiated only above 30 MHz.

Figure 2 shows the FCC limits. All of the figures with commercial radiated limits have been scaled to a measurement distance of 3 meters to simplify comparisons. Actual measurement distances may vary according to the appropriate specification.

As can be seen from Figure 2, the Class A limits are less stringent than the Class B limits. The paperwork is also less stringent with Class A, which requires only verification by testing. Class B equipment, on the other hand, must be certified, which requires both testing and the filing of an application with the FCC. Since the Class B limits are more stringent, they must be applied unless it can be shown that the equipment will not be used in the home.

Presently, Part 15J requirements apply primarily to home and business computers. Exceptions are given for test equipment, appliances, industrial controls, and transportation equipment. These exemptions may change in the future, particularly if interference problems from these sources become widespread.

Additional high visibility FCC requirements for EMI include Part 18 (industrial, scientific, and medical equipment) and Part 68 (equipment connected to the telephone network). Part 18, like Part 15, is an emission specification with an objective of preventing interference to authorized radio services. Part 68, on the other hand, focuses on compatibility with the telephone network.

Although the FCC Part 15 regulations address only emissions, the laws provide for susceptibility testing as well. Presently, the FCC has not mandated susceptibility limits for commercial equipment, preferring to leave that to the individual manufacturers.

VDE 0871 and 0875

In Europe, most countries have their own agencies and regulations regarding EMI. There have been broad efforts at coordinating those limits, however, particularly within the European Economic Community. One of the more strict, and widely accepted, set of standards is that of West Germany, known as the VDE regulations.

Unlike the FCC, the VDE (Verband Deutscher Electrotechniker, or Association of German Electrical Engineers) does not have regulatory powers. Rather, the VDE issues standards, much like the IEEE (Institute of Electronic and Electrical Engineers) or SAE (Society of Automotive Engineers) in the United States. Those standards are subsequently incorporated into the German laws by decree.

The legal responsibility lies with the Deutsch Bundespost, or German Post Office, and is administered through the FTZ, or Central Telecommunications Office. The VDE does, however, retain the testing responsibility, which is similar to that of the FCC in the United States.

Two VDE regulations for EMI are VDE 0871 and 0875. VDE 0871 applies to computer equipment and parallels FCC Part 15J. VDE 0875 applies to household appliances, electrical appliances and tools, and fluorescent lights — all areas currently unregulated in the United States. Figure 3 summarizes the VDE 0871 levels. Compared to the FCC limits, the VDE requirements are broader in scope and frequency range than the FCC requirements. In the frequency areas of overlap, however, they are quite similar.

Equipment subject to the VDE 0871 requirements is divided into three classes. Class A applies to commercial equipment, Class B to residential, and Class C is a special site certification provision. Since the Class A limits are less stringent than the Class B limits, the VDE is more particular about the testing and paperwork. This is opposite the FCC approach. In any event, products that meet the less stringent Class A limits must be tested by the VDE or at a VDE-approved lab, and additional paperwork must be submitted.

Products that meet the more stringent Class B limits can be self-certified. Thus, many manufacturers prefer to suppress their emissions to Class B limits, even if their equipment is to be used only in the commercial environment, saving test time and money and reducing the time to market. The best of both worlds, for the United States and European markets, is to be eligible for Class A (easier to test in the United States) and to meet Class B (easier to test and market in Europe).

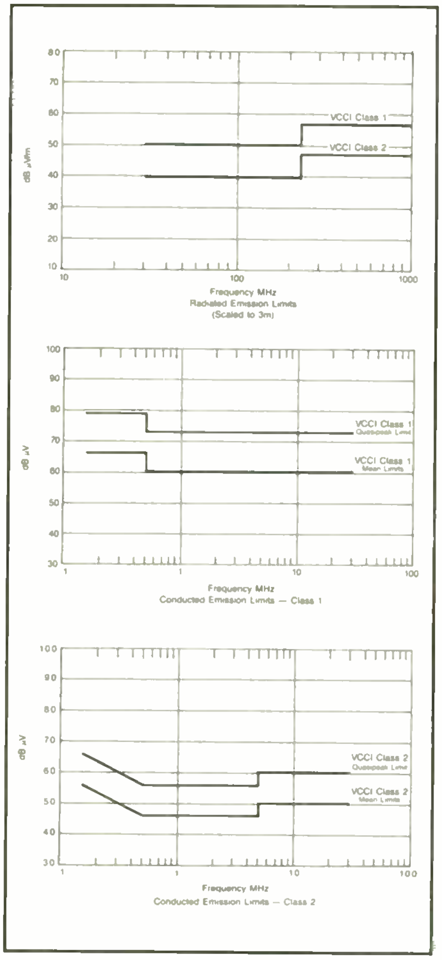

Figure 4. VCCI limits for radiated and conducted emissions.

Like the FCC, the VDE 0871 requirements are emission-related. The radiated electric field limits are similar, with some small differences due to different television frequencies. The conducted limits extend to lower frequencies with the VDE limits. In addition, the VDE also tests for magnetic field emissions, which is not required by the FCC.

The VDE is also responsible for product safety in West Germany. These limits are also quite stringent, and in general, products that meet North American standards may not be accepted as safe in Germany or the rest of Europe.

VCCI

In March 1986, Japan instituted a set of voluntary test requirements for limiting emissions from computer equipment. These requirements were established by the VCCI (Voluntary Control Council for Interference by Data Processing Equipment). Although still voluntary, these requirements are expected to become mandatory in the future.

Like the FCC and VDE, the VCCI limits are subdivided into two classes. Class 1 equipment applies to ITE (Information Technology Equipment) that is designed for use in industrial or commercial environments, and Class 2 equipment is equipment designed for residential use.

The VCCI limits are shown in Figure 4, and apply to equipment manufactured after December 1988. The VCCI had applied a phased application of the limits, giving manufacturers a cushion during the transition to the requirements.

Summary

Although there are several different EMI regulations that affect electronic equipment, the general goal is the same — to assure successful operation of the equipment itself without causing upsets to nearby equipment.

In the military environment, this generally means testing for both emissions and susceptibility, to ensure operating margins within a system. In the commercial environment, this usually means only testing for emissions, against limits that ensure that nearby television and radio receivers will not be adversely affected. rf

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About the Author

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Bridged-Tee Delay Equalizers— A Computer-Aided Realization

By Robert C. Kane
Motorola, Inc.

With the many types of signal transmission that require accurate recovery of information, distortion of any kind is a problem. RF engineers are increasingly asked to reduce distortion in systems as performance demands escalate. One important problem area is time distortion, or group delay, where a signal experiences unequal transit times for different frequencies within a system passband. This article describes a computer-assisted design method for group delay equalizers, which reduce the delay distortion in RF systems.

Group delay distortion associated with a signal, or group of frequencies, passing through a network or circuit results in a degradation of system performance. In some instances the degradation will be insignificant, such as when a relatively narrow-band signal is passed through a network with a wide bandwidth response. An example would be an 18 GHz carrier modulated to achieve a maximum 6 MHz bandwidth passing through a 100 MHz bandwidth preselector filter. In the region close to the center frequency of the filter, the group delay distortion contribution is insignificant and can generally be ignored.

However, if the modulation of that same carrier were to increase to 30 MHz, the group delay distortion becomes considerable as the dispersion of the signal through the network increases to those components of the signal farther away from the center frequency, as shown in Figure 1.

In a system where these performance degradations cannot be tolerated, delay equalizers are employed to generate a delay function which is opposite to that of the delay response of the un-equalized system. For a properly designed equalizer, the cascaded responses will yield a constant propagation delay over the frequency range of interest (Figure 2).

It should be noted that in most instances it is not necessary or desirable to equalize the delay over an entire passband. Rather, only the narrower region of the occupied bandwidth need be considered. In the previous example, equalization over the 30 MHz modulation bandwidth would be adequate. Equalizing over the entire passband of the filter would be unnecessary and would increase the complexity and cost of the equalizing network.

The purpose of this paper is to present a systematic approach to the realization of bridged-tee delay equalizers of the order 1 through 8. The development leads to a computer program which, when implemented, eliminates the previously used trial-and-error approximation iterations.

The Design Method

The development by Phillips (1) is used as the basis for the work which follows. In those papers a procedure was presented whereby standardized delay curves may be plotted utilizing the methods of Skwirzynski (2). It was left to the circuit designer to match a delay function to be equalized by iterating various equalizer responses onto the standardized curves until a suitable correlation was achieved.

In this paper, the development effort is reduced to a software routine which performs the iteration over a specified range of parameters and searches for the most appropriate equalizer.

Placing the equalizer development into a software routine requires that the delay response to be equalized must be represented by a suitable function. The Newton interpolation method using Chebyshev nodes is utilized for this purpose.

For n nodes:

$$x_i = 1/2(a + b) + 1/2(b - a)\cos[(2i-1)\pi/2n] \quad 1 \leqslant i \leqslant n$$

For our application the interval always begins at 0.0, which allows us to simplify the above equation to:

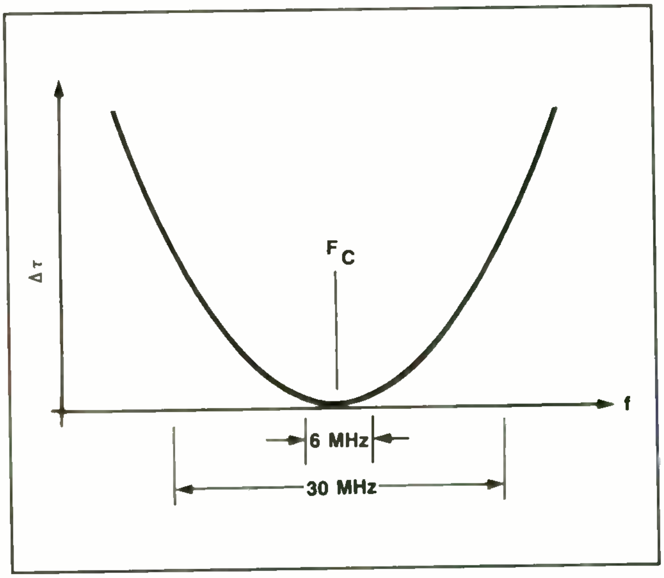

Figure 1. Delay effects versus bandwidth.

$$x_j = b/2 + b/2\{\cos[(2i-1)\pi/2n]\}$$

$$x_j = b/2\{1 + \cos[(2i-1)\pi/2n]\}$$

where b is the upper end of the interval corresponding to the upper limit of the range of frequencies to be equalized.

From the x_j, corresponding values of delay are determined from the filter data.

$x_j = \Delta f \qquad x_1, x_2, x_3, \ldots\ x_n$

$f(x_j) = t_d \qquad f(x_1),\ f(x_2),\ f(x_3),\ \ldots\ f(x_n)$

The polynomial to be generated takes the form:

$$P(x) = a_1 + a_2(x - x_1) + a_3[(x - x_1)(x - x_2)] + a_4[(x - x_1)(x - x_2)\ \ (x - x_3)\ \ \ldots a_n[(x - x_1)(x - x_2)\ldots(x - x_n)]$$

$$P(x) = a_1 + (x - x_1)[a_2 + (x - x_2)[a_3 + (x - x^3)[\ldots]]]]]]$$

The polynomial coefficients are determined as follows:

$$a_1 = f(x_1)$$

$$f(x_1) = a_1$$

$$a_2 = [f(x_2)\ f(x_1)]/(x_2 - x_1)$$

$$f(x_2) = a_1 + a_2(x_2 - x_1)$$

$$a_3 = [f(x_3) - f(x_1) - (f(x_2 - f(x_1))(x_3 - x_1)]/[(x_3 - x_1)(x_3 - x_2)]$$

$$f(x_3) = a_1 + a_2(x_3 - x_1) + a_3(x_3 - x_1)(x_3 - x_2)$$

$$\begin{aligned} a_4 = \{&f(x_4) - f(x_1) - [(f(x_2) - f(x_1))(x_4 - x_1)/(x_2 - x_1)] \\ &- [f(x_3) - f(x_1) - (f(x_2) - f(x_1))(x_3 - x_1)/((x_3 - x_1)(x_3 - x_2))] \\ &(x_4 - x_1)(x_4 - x_2)\}/(x_4 - x_3)(x_4 - x_2)(x_4 - x_1) \end{aligned}$$

$$\begin{aligned} f(x_4) = &a_1 + a_2(x_4 - x_1) + a_3(x_4 - x_1)(x_4 - x_2) \\ &+ a_4(x_4 - x_3)(x_4 - x_2)(x_4 - x_1) \end{aligned}$$

Which can be represented as,

$$f[x_1 \ldots x_k] = \{\ f[x_2 \ldots x_k] - f[x_1 \ldots\ldots x_{k-1}]\ \}\ /\ (x_k - x_1)$$

Therefore,

$$f[x_j] = f(x_j)$$

$$f[x_j, x_{j+1}] = \{\ f[x_{j+1}] - f[x_j]\ \}\ /\ (x_{j+1} - x_j)$$

$$f[x_j, x_{j+1}, x_{j+2}] = \{\ f[x_{j+1}, x_{j+2}] - f[x_j, x_{j+1}]\ \}\ /\ (x_{j+2} - x_j)$$

When placed in the form:

$$\begin{array}{lllll} x_1 & f[x_1] & & & \\ & & f[x_1,x_2] & & \\ x_2 & f[x_2] & & f[x_1,x_2,x_3] & \\ & & f[x_2,x_3] & & f[x_1,x_2,x_3,x_4] \\ x_3 & f[x_3] & & f[x_2,x_3,x_4] & \\ & & f[x_3,x_4] & & \\ x_4 & f[x_4] & & & \end{array}$$

the upper diagonal yields the desired polynomial coefficients.

$$P(x) = f[x_1] + f[x_1,x_2](x - x_1) + f[x_1,x_2,x_3](x - x_1)(x - x_2)\ldots$$

Recall that this function is unique for each value of t_{te} (maximum equalizer delay) used. Therefore, the polynomial is determined each time the program iterates on t_{te}.

Before the delay function can be compared to the standardized functions, the polynomials of the standardized functions must be generated. This is a one time event and the results will vary only by a scaling factor.

For the one section equalizer,

$$t_{norm} = 1\ /\ (1 + e\Omega^2) \qquad e = \pi^2/3 \text{ the scaling factor}$$

For the n-tuple section equalizer,

$$\begin{aligned} t_{norm} = 1/M \Big\{ &[1/\ ((\Omega' + L_1)^2 + 1) + 1/((\Omega' - L_1)^2 + 1)] \\ &+ K_2[1/\ ((K_2\Omega' + L_2)^2 + 1) + 1/((K_2\Omega' - L_2)^2 + 1)] \\ &+ K_3[1/((K_3\Omega' + L_3)^2 + 1) + 1/\ ((K_3\Omega' - L_3)^2 + 1)] \\ &+ K_4[1/((K_4\Omega' + L_4)^2 + 1) + 1/((K_4\Omega' - L_4) + 1)] \\ &+ K_0\ /\ ((K_0\Omega')^2 + 1) \Big\} \end{aligned}$$

Where,

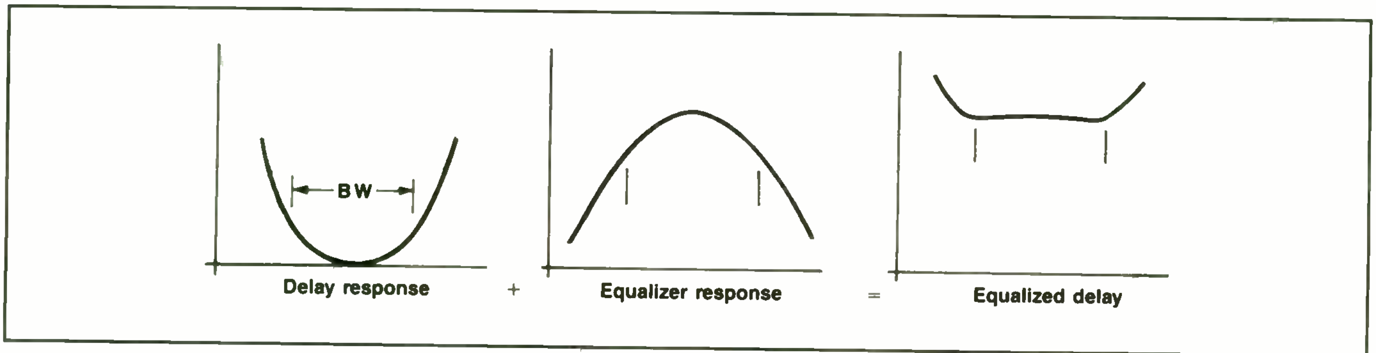

Figure 2. Delay equalizer operation.

$$M = 2/(L_1^2 + 1) + 2K_2^2/(L_2^2 + 1) + 2K_3^2/(L_3^2 + 1) + 2K_4^2/(L_4^2 + 1) + K_0$$

and,

$L_j = A_jP_j \quad K_j = B_jH_j$ (See Ref. #1 for table of values)

Also,

$$\Omega' = (\pi/\sqrt{3})(\Omega/M)$$

In order to preserve the correlation, when integrating to find the closest fit, remember to scale the normalized frequency by the same factor.

The process now evolves as follows:

1. Select data nodes by Chebyshev node algorithm.
2. Take data at these nodes.
3. Select a maximum filter equalizer delay t_{te}.
4. Normalize frequency to map onto the standard delay functions.

$$\Omega_j = x_j t_{te}$$

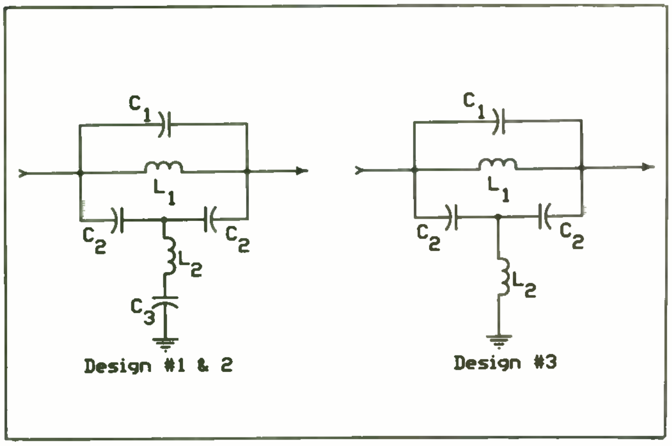

Figure 3. Bridged-Tee equalizer configurations.

5. Construct the function polynomial.
6. Integrate to determine area between function polynomial and standard delay functions.
7. Iterate for n section ($1 \leq n \leq 8$).
8. Iterate over the range of allowable equalizer delay.
9. Select the appropriate n-section equalizer which yields minimum error.
10. Repeat for the modified delay functions for a specific nth-order equalizer and select best fit.
11. Realize n-section Bridged-Tee cascaded networks.

The integration is accomplished numerically by the Trapezoidal method with 1000 sub-intervals in order to achieve a significant degree of accuracy when comparing values of the function associated with a given t_{te} to those of the adjacent standard delay functions.

Software Implementation

Note that the designer is given the option of selecting from a number of solutions. These solutions are identified as Designs 1, 2 and 3, which take the configurations of Figure 3. The difference between Design 1 and 2 lies only in the resultant ratio of capacitance values.

The net deviation for a given t_{te} and equalizer standard function (E_n) is:

$$\sum_1 | P_j - E_j |$$

Upon completion of iteration of n and the desired allowable range of t_{te}, the best fit (least area of deviation) determines the n and t_{te} to be used in realizing the actual circuit. A printout of a design example is shown in Figure 4.

The program will prompt the designer for inputs such as center frequency, bandwidth, allowable delay range and terminating impedances. Since the delay is assumed to be symmetric about the center frequency, the relative fit can be determined over either the upper or lower half-band. The printout of the example demonstrates this with a listing of the

```
NETWORK TERMINATING IMPEDANCE =  75 OHMS

DELAY RANGE OF EQUALIZER IS  40nSEC. TO  80nSEC.

DELAY RESPONSE CENTER FREQ. IS  140MHz

EQUALIZES BANDWIDTH  60MHz

DATA POINT # 1 @  140.2278MHz

DATA POINT # 2 @  142.0096MHz

DATA POINT # 3 @  145.3581MHz

DATA POINT # 4 @  149.8696MHz

DATA POINT # 5 @  155MHz

DATA POINT # 6 @  160.1303MHz

DATA POINT # 7 @  164.6418MHz

DATA POINT # 8 @  167.9903MHz

DATA POINT # 9 @  169.7721MHz


 80 nSEC @  140MHz
 80.1 nSEC @  140.2278 MHz
 80.2 nSEC @  142.0096 MHz
 80.4 nSEC @  145.3581 MHz
 80.8 nSEC @  149.8696 MHz
 81.6 nSEC @  155 MHz
 83.2 nSEC @  160.1303 MHz
 86.4 nSEC @  164.6418 MHz
 92.8 nSEC @  167.9903 MHz
 105.6 nSEC @  169.7721 MHz

 3SECTION EQUALIZER REQUIRED

 48 nSEC.  MAX. EQUALIZER DELAY
 .533948740317 RELATIVE MATCH FACTOR
REVISED FORM # 1 IS BEST FIT
 .483483677352 RELATIVE MATCH FACTOR
```

```
*************EQUALIZER SECTION # 1 ***************

DELAY CENTER FREQ. =  163.273688656MHz

EQUALIZER DESIGN #1  C2=C3

C1 =  38.6pF   C2 =  6pF   C3 =  6pF
L1 =  .022uH   L2 =  .234uH

EQUALIZER DESIGN #2   C1=C2

C1 =  27.8pF   C2 =  27.8pF   C3 =  4.3pF
L1 =  .022uH   L2 =  .234uH

EQUALIZER DESIGN #3    C3=INF. CAP.

C1 =  40.6pF   C2 =  2pF
L1 =  .022uH   L2 =  .234uH
*************EQUALIZER SECTION # 2 ***************

DELAY CENTER FREQ. =  116.726311344MHz

EQUALIZER DESIGN #1  C2=C3

C1 =  35.7pF   C2 =  11.8pF   C3 =  11.8pF
L1 =  .044uH   L2 =  .234uH

EQUALIZER DESIGN #2   C1=C2

C1 =  27.8pF   C2 =  27.8pF   C3 =  9.2pF
L1 =  .044uH   L2 =  .234uH

EQUALIZER DESIGN #3    C3=INF. CAP.

C1 =  39.7pF   C2 =  3.9pF
L1 =  .044uH   L2 =  .234uH
*************EQUALIZER SECTION # 3 ****************

DELAY CENTER FREQ. =  140MHz

EQUALIZER DESIGN #1  C2=C3

C1 =  31.5pF   C2 =  9.4pF   C3 =  9.4pF
L1 =  .035uH   L2 =  .204uH

EQUALIZER DESIGN #2   C1=C2

C1 =  24.2pF   C2 =  24.2pF   C3 =  7.2pF
L1 =  .035uH   L2 =  .204uH

EQUALIZER DESIGN #3    C3=INF. CAP.

C1 =  34.7pF   C2 =  3.1pF
L1 =  .035uH   L2 =  .204uH
```

Figure 4(a). Design example printout.

calculated Chebyshev nodes for a 60 MHz bandwidth centered at 140 MHz. Notice that only the nodes for the upper half-band are needed.

The length of the program precludes its publication here. The program is available on disk from the RF Design Software Service (see page 78 for details). A printout of the program listing can be obtained by sending a self-addressed legal-size envelope with 45 cents postage to RF Design. Please note that the program is written in HP BASIC. The listing is only available in this form. The disk version will have the HP BASIC version on an MS-DOS readable disk.

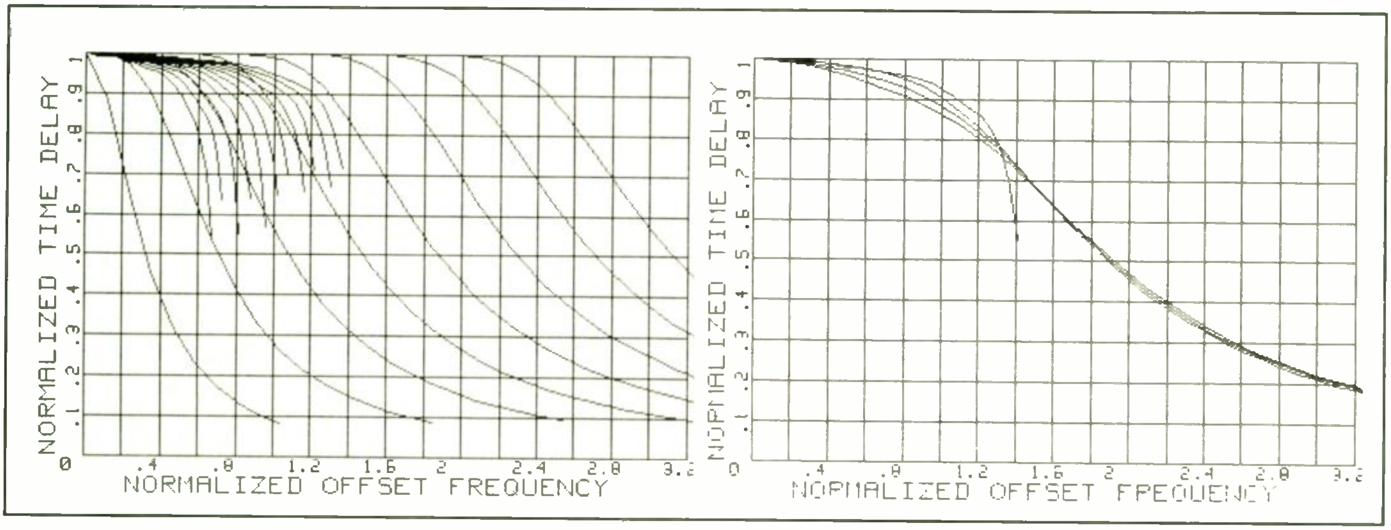

Figure 4(b). "Best fit" iterations (left) and equalized results (right).

About the Author

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References

1. Richard R. Phillips, "Curve-Matching Method Specs Delay Equalizers," Parts I and II, *Microwaves,* September 1980, pp. 67-78, and October 1980, pp. 65-71.
2. J.K. Skwirzynski, *Design Theory and Data for Electrical Filters,* Van Nostrand, 1965, pp. 188-243.

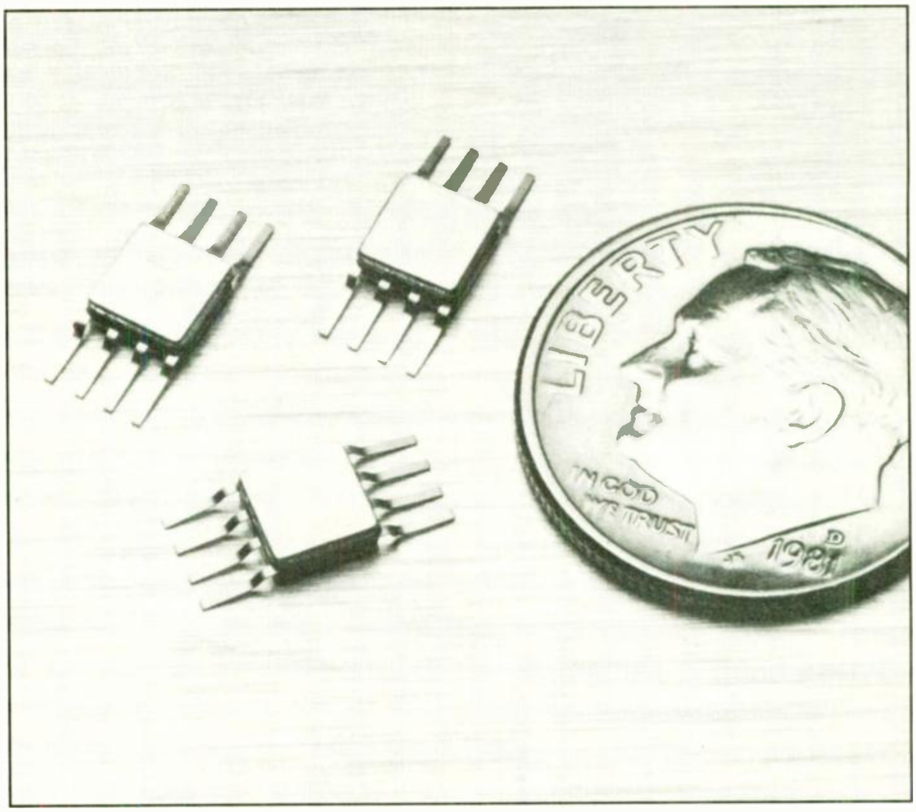
Avantek Introduces Si MMIC Active Mixer/Amplifiers

Designated IAM-81018 and IAM-82018, these active mixers operate with RF input signals from 5 MHz to 5 GHz with IF outputs from DC to 1 GHz and 2 GHz, respectively. The 81018 provides typical RF-to-IF gain of 8 dB, and operates with a -5 dBm LO level. The 82018 provides typical RF-to-IF gain of 15 dB, operating from a 0 dBm LO level. Both of these parts are inherently insensitive to load characteristics.

Specifications include IF output at 1 dB gain compression of -6 dBm typ, third-order intercept point of 3 dBm, and SSB noise figure of 15 dB. VSWR is 1.5:1 between 0.05 and 5 GHz for the RF and LO ports.

The units are packaged in a hermetic 180-mil surface-mount package with gold-plated leads, and are fully compatible with 50-ohm microstrip systems. They are built with a fully isolated interdigitated bipolar process which results in f_T of up to 15 GHz and f_{max} of 25 GHz.

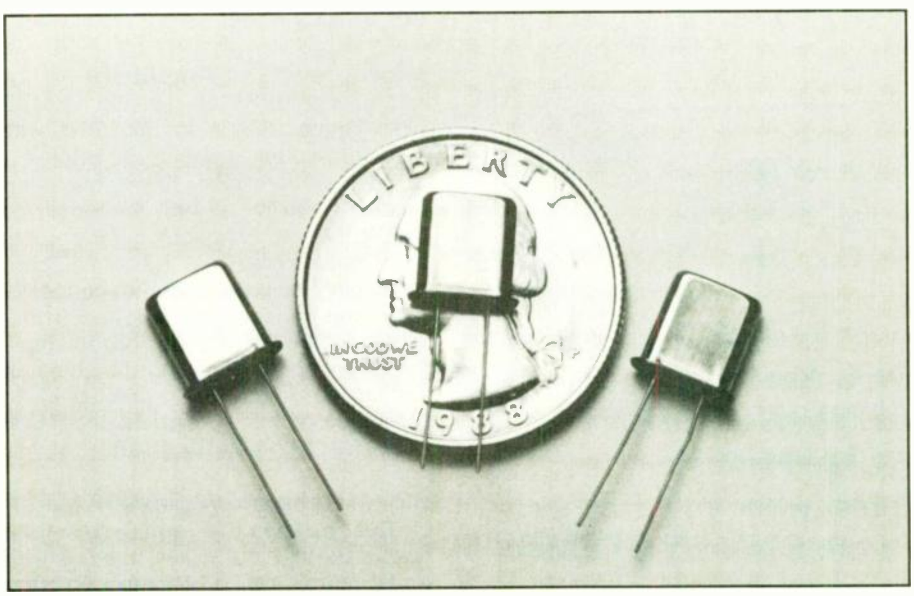
The Avantek IAM-81018 is priced in the $16 range, while the IAM-82018 is priced in the $25 range — both in 100-piece quantities. **Avantek, Inc., Santa Clara, CA. INFO/CARD #200.**



Miniature Quartz Crystals From Savoy

The HC 45/U and HC 80 high frequency crystals are available for frequencies from 100 to 250 MHz with maximum frequency tolerance of ± 30 ppm. Temperature stability for a 250 MHz unit is ±20 ppm from -10 degrees C to 60 degrees C. Equivalent series resistance is 60 ohms max for the 100 to 150.99 MHz range, 80 ohms for 151 to 200.99 MHz range and 120 ohms max for the 201 to 250 MHz range. Modes of oscillation for the above frequency breaks are at 5th, 7th and 9th overtones, respectively.

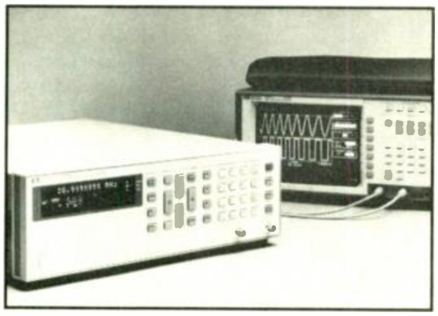
Aging is 5 ppm/year and drive level is set at 1 mW. The shunt capacitance is 7 pF while the load capacitance is determined by the particular requirement. The HC 45/U package features maximum dimensions of 0.319 in. X 0.272 in. X 0.126 in., with a tolerance of ±0.0005 percent. For improved aging characteristics, a cold weld sealing method is used for the HC 45/U and resistance weld is used for HC 80.



Custom specifications are available from 1 MHz, and overall pricing for the crystals is determined by customer specifications and quantity. **Savoy Electronics, Inc., Ft. Lauderdale, FL. Please circle INFO/CARD #199.**

Synthesized Function/Sweep Generator

The HP 3324A is a synthesized function/sweep generator that can be used as an alternative to standard function generators. It offers a frequency accuracy of 5 ppm, -55 dBc spurious distortion and is characterized by -50 dBc phase noise. The instrument generates five waveforms: sine, square, triangle, negative and positive ramps with frequency coverage from 1 mHz to 21 MHz (sinewave). A 1 mHz to 60 MHz TTL clock is incorporated in the generator for use in operations such as clock generation for fast A/D or D/A converters.



Amplitude (10 V p-p) and DC offset (±4.5 V) can be set with 4-digit resolution. A high-voltage option, which increases the output signal by a factor of four, is also available. When the waveforms are turned off, the instrument can be used as a DC source. The basic unit is priced at $3,500. **Hewlett-Packard Company, Palo Alto, CA. Please circle INFO/CARD #198.**

Miniature Power Amplifiers

Model AMPS1150 exhibits 13 dB typical gain from 1 to 500 MHz with minimum output power of 30 dBm at 1 dB gain compression. The 2-tone intermod is typically -38 dBc referenced to the PEP level. Efficiency is 20 percent and input VSWR is 1.7:1. Available packaging includes solder-in, connectorized and finned-heatsink styles. The 1 watt, 1 to 500 MHz connectorized amplifier is priced at $275 and the solder-in version is $120 when purchased in quantities of 1 to 4. **Acrian, Inc., San Jose, CA. INFO/CARD #197.**

Signal Generator/Deviation Meter

This synthesized signal generator covers the 0.01 to 550 MHz range with an RF output range of 13 to -127 dBm with output accuracy of ± 1.5 dB and full band flatness of ± 1.0 dB. Standard features include software-controlled self-calibration for frequency and amplitude, GPIB interface, and external clock input/output. Model 2407 is priced at $4595. **Wavetek RF Products, Inc., Indianapolis, IN. INFO/CARD #196.**

Voltage Controlled Oscillator

EMF Systems introduces a VCO with a 1.5 to 2.7 GHz frequency range in four bands. Power is 10 dBm and harmonics are measured at 15 dBc. At 150 kHz frequency offset, the single-sideband phase noise is measured at 120 dBc/Hz. **EMF Systems, Inc., State College, PA. INFO/CARD #195.**

RF Power MOSFETs

M/A-COM PHI introduces the LF Series of N-channel enhancement mode RF power MOSFETs. The LF2802A features DMOS structure, a frequency range of 500 MHz to 1400 MHz, and 2 watts Class A output. Gain is 10 dB at 1 GHz, VSWR is 10:1, and forward transconductance g_m is 40 mhos. Other members of the series feature different power levels. **M/A-COM PHI, Inc., Torrance, CA. INFO/CARD #194.**

Receive-Only Antennas

Mark Antennas Division introduces a line of receive-only antennas in both parabolic grid and solid configuration. Typical applications include ITFS, MMDS and MDS bands. **Radiation Systems, Inc., Mark Antennas Div., Des Plaines, IL. INFO/CARD #193.**

150 MHz Op Amp

CLC500 is an operational amplifier that settles to 0.01 percent in 25 ns max. Since the device uses a current-feedback architecture, the settling performance is independent of gain setting. It offers 12-bit performance and features output voltage clamping which allows designers to set the maximum positive and negative output voltage swings to protect sensitive A/D systems from overload and saturation. At 20 MHz and 2 V p-p, the second harmonic distortion is typically -50 dBc while the third harmonic distortion is -60 dBc typ. Available packaging includes 14-pin plastic DIP and hermetic sidebrazed ceramic 14-pin DIP. In 100-piece quantity, price ranges from $17.35 to $43.30, each depending on packaging. **Comlinear Corp., Ft. Collins, CO. INFO/CARD #192.**

DDS With Decimal Resolution

The DRFS-VR1070 is a direct digital synthesizer that provides exact decimal frequency resolution. With the on-board or external 10 MHz reference, the device will synthesize frequencies from DC to 3.3 MHz with 1 Hz resolution. Output spurious levels are below -70 dBc and settling time and agility between any output frequency is 500 ns max. **Digital RF Solutions Corp., Santa Clara, CA. INFO/CARD #191.**

Tubular Filters

A line of "In-a-Cable" lowpass, highpass and bandpass filters is available from Micro-Coax. The design permits the filter element to be imbedded within the cable, eliminating up to four previously required connectors. Lowpass filters are available in cutoff frequencies from 0.25 to 26.5 GHz in cable sizes of 0.085 in., 0.141 in., and 0.250 in. The bandpass filters provide center frequencies from 1 to 23 GHz and come in a cable diameter of 0.141 in. Cutoff frequencies from 2 to 5 GHz in 0.141 in. diameter are available for the highpass configuration. **Micro-Coax Components, Inc., Collegeville, PA. Please circle INFO/CARD #190.**

Gold Metallized Power FETs

Semetex introduces the ST1016 silicon gold-metallized RF power GigaFET that features push-pull configurations to 2 GHz. The devices are tested at 500 MHz, 10 watts and 13 dB gain. **Semetex Corp., Newbury Park, CA. Please circle INFO/CARD #189.**

Second Generation Tuning Stick

The second generation of Tuning Stick kits from ATC includes all in-between values from 0.1 pF to 1000 pF. Featured are 27 different values which contain ATC 100 Series Superchip^{R} radial wire leaded capacitors labeled with their specific values. Each capacitor is attached permanently to a nonconductive holder. **American Technical Ceramics Corp., Huntington Station, NY. Please circle INFO/CARD #188.**

Instrument Amplifier

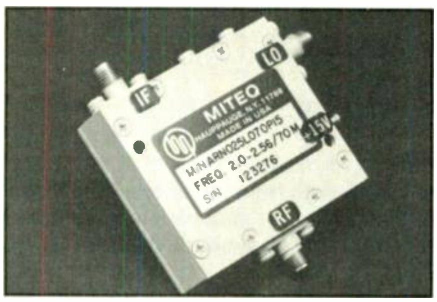
Marconi Instruments introduces a 10 MHz to 4.2 GHz amplifier which boosts the input sensitivity and tracking generator output of the 2380 Series of spectrum analyzers. When used with the 2380 Series, the amplifier delivers a gain of at least 25 dB. As a preamplifier, the input sensitivity of the spectrum analyzer is normally increased by 25 dB, allowing the instrument to view low level signals. The tracking generator output is boosted to 15 dBm providing a range of output levels from -20 dBm to 15 dBm. The amplifier can also be used to boost the output of other instruments. **Marconi Instruments, Inc., Allendale, NJ. INFO/CARD #187.**

DIP VCO

Model VC-373 is a non-crystal-controlled VCO providing a TTL output at any frequency from 1 MHz to 90 MHz; HCMOS output is available up to 40 MHz. Linearity and deviation are ± 10 percent and Class B screening to MIL-O-55310 is available. In the 25-piece quantity, the VCO costs $84. **Vectron Laboratories, Inc., Norwalk, CT. Please circle INFO/CARD #186.**

Image Reject Low-Noise Front-End

Miteq's ARN0251070P15 image reject low-noise front-end has an RF/LO frequency range of 2 to 2.5 GHz with an IF



frequency of 70 MHz. Gain is typically 30 dB and noise figure is less than 1.2 dB. Image rejection is greater than 20 dB and the output at the 1 dB compression point is 12 dBm. Bandwidths are 10 to 35 percent of center frequency in the 0.4 to 10 GHz range. A switchable sideband option is also available. **Miteq, Hauppague, NY. INFO/CARD #185.**

30 A MOSFET Hybrid

The DE-375X4 Macho MOS is a 1000 V, 30 A, MOSFET hybrid that operates at 30 kW at frequencies in excess of 1 MHz. Switching time is less than 10 ns with 400 kW peak power and 400 A peak current. Applications for the device include radar, sonar, HF communications, lasers, EMP and power conversions. In evaluation-size quantities, the price is $550. **Directed Energy, Inc., Fort Collins, CO. INFO/CARD #184.**

Frequency Agile Signal Simulator

The HP 8791 Model 10 hardware platform provides an agile carrier that switches in less than 250 ns over a 10 MHz to 3 GHz bandwidth. Frequency resolution is 0.125 Hz and the carrier has a direct-digital synthesis modulation capability with a 40 MHz instantaneous bandwidth, allowing for spread-spectrum formats such as chirps, Barker codes, maximal-length sequences, QAM, and FSK. For applications requiring carriers above 3 GHz, band-oriented up-converters are available to 18 GHz, providing 2 GHz of agile bandwidth. Also available is the HP 8791 Model 100 precision signal generator ID, HP 8791 Model 200 radar simulator ID and HP 11776A Option K10 waveform generation software. **Hewlett-Packard Company, Palo Alto, CA. INFO/CARD#183.**

SiO_2 Cables With QD Connectors

Kaman introduces SiO_2 connectors with quick disconnect (QD) connectors which mate with standard female connectors. Connector insertion loss per connector pair is 0.14 dB and VSWR is 1.2 at 2 GHz. **Kaman Instrumentation Corp., Colorado Springs, CO. Please circle INFO/CARD #182.**

Receiver Multicoupler System

This receiver multicoupler system has a frequency range of 5 kHz to 900 MHz with 0 dB gain from input to output. Four separate receiver multicouplers are used to cover individual bands from 5 kHz to 2 MHz, 2 MHz to 30 MHz, 30 MHz to 200 MHz, and 200 MHz to 900 MHz. In the HF and VHF stages, reverse isolation is 120 dB, third-order intercept point is 53 dBm min, and output-to-output isolation is 40 dB. The VLF/LF and UHF receiver multicouplers have reverse isolation of 90 dB and output-to-output isolation of 25 dB. **Wi-Comm Electronics, Inc., Massena, NY. INFO/CARD #181.**

HF Tuner

The Steinbrecher Model 12102A HF tuner provides 96 dB of dynamic range in a 2 MHz bandwidth with noise figure of 12 dB. The half-rack-wide unit allows side-by-side mounting in a standard 19 in. RETMA rack. When connected to an appropriate analog-to-digital converter, the tuner supports 16 bits of resolution. **Steinbrecher Corp., Woburn, MA. INFO/CARD #180.**

GaAs MMIC Attenuator

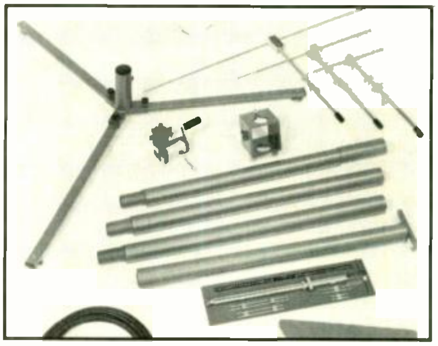
This line of GaAs linear and nonlinear MMIC attenuators covers the DC to 6 GHz range. Model 2201-23-022 has an attenuation range of 10 dB with linearity of 2.15 dB/V and accuracy of ±0.2 dB. The input power at 1 dB attenuation change is 2 dBm and maximum VSWR in and out is 1.7:1. When purchased in small quantities, price is $120 each. **Midisco, Commack, NY. Please circle INFO/CARD #179.**

Rubidium Frequency Standard

The Austron Model 2112 provides TTL outputs of 1 MHz, 5 MHz, 10 MHz or 100 kHz; sinewave outputs of 1 MHz, 5 MHz and 10 MHz; and a 1 pps clock output that can be synchronized externally. Options include dual reference input, IEEE-488, and an internal battery backup. **Austron, Inc., Austin, TX. INFO/CARD #178.**

EMI/Field Strength Antennas

Schaffner EMC introduces a line of antennas manufactured by Schwarzbeck Electronik AG. The VHA-9103, a dipole for use from 30 to 300 MHz, and the UHA-9105, a dipole with a 300 to 1000 MHz range, are available in two



versions. The precision version, with a compensating attenuator, produces an output with a flatness of ±0.1 dB. The standard version is provided with a calibration curve which permits manual correction readings to similar accuracies. The precision dipoles cost $2,200 each and the standard units are priced at $1,350. **Schaffner EMC Inc., Union, NJ. INFO/CARD #176.**

SMT Beam Lead Diode Packages

These surface mount beam lead diode packages are available in 0.050 in., 0.070 in., 0.0850 in., and 0.100 in. octagon sizes. Designed for mixer and detector applications up to 18 GHz, these ceramic packages are available in two-, three- or four-lead versions with 0.010 in. and 0.020 in. standard thicknesses of 96 percent alumina ceramic. The packages are screened per MIL-STD-105D, MIL-STD-883, and MIL-STD-750 requirements and are plated with 50 micro-in. nickel and 50 micro-in. gold min. **Cabot Ceramics, Greenville, RI. INFO/CARD #175.**

rf software

Waveform Processing Software

W.A.V.E. is a data acquisition and analysis tool that can directly create, acquire, and manipulate waveforms. Features include macro-capability, workspace and macro editors, a waveform element editor, and Lotus 1-2-3 compatibility. The software supports various data types including waveforms, vectors, matrices, scalars and strings. Rectangular plots, Smith charts, polar plots, contour charts, and 3-D mesh plots are provided. W.A.V.E. is designed for use on IBM-PC/XT/AT and IBM PS/2 computers and compatibles. It is priced at $495. **Vespine Corp., Urbana, IL. Please circle INFO/CARD #230.**

EMC Database and Engineering Solver

PT Express is a memory resident program that features conversions, equations, math and tables tailored towards the EMC and RF engineer. The user can employ the standard templates or create specific ones for the operations mentioned. For equations, the software features Ohm's law, reactance, resonance, and VSWR. The Math section provides a full screen editor-type calculator/plotter, and the tables featured include FCC Part 15-J, Capacitor Self-Resonance, E-Field to Power Density, and Skin Depth. The software is priced at $99.95. **Liberty Labs, Inc., Cedar Rapids, IA. Please circle INFO/CARD #229.**

Component Matching Software

Component Matching Software matches measured component characteristics by specified tolerances, and produces component groupings by the maximum number of matched sets. It operates from a table of measured data which lists measurement points and matching tolerances by component identification number. A typical application is for providing cable in matched pairs. **Test Quality Company, Santa Clara, CA. Circle INFO/CARD #228.**

Radar-Jammer Simulation System

Radar Vulnerability Assessment System (RVAS) is a software system of VAX computers which models radar receiver systems and evaluates their response to jamming. The software provides a menu-oriented interface through which the user can construct block diagram circuit models of radar systems and combine them with jammers and targets. **Research Associates of Syracuse, Inc., North Syracuse, NY. INFO/CARD #227.**

Software Data Sheets

DGS Associates announces the release of data sheets describing the advanced features and benefits of S/FILSYN™ —a microwave filter design software program. As described in the data sheets, the software is an interactive filter design, synthesis and analysis program which is offered in modules and comes with a number of utilities. Featured in the literature are common characteristics, physical implementations, general features, complementary programs and hardware requirements. **DGS Associates, Inc., Santa Clara, CA. INFO/CARD #216.**

Radiation Meters Brochure

This publication features RF and microwave field strength meters for measurement of non-ionizing radiation. Incorporated are microwave oven leakage meters, isotropic field strength meters, and meters for measuring electric and magnetic fields near VDTs and electric power lines. **Holaday Industries, Inc., Eden Prairie, MN. Please circle INFO/CARD #220.**

Quartz Crystals and Oscillators

Piezo Crystal introduces its catalog covering quartz crystals and crystal oscillators for commercial, military and spacecraft applications. Technical and product information on quartz crystals and quartz oscillators up to 600 MHz are featured. **Piezo Crystal Company, Carlisle, PA. INFO/CARD #219.**

RCS Target Mounting Systems Brochure

Described in this brochure is Scientific-Atlanta's line of Series 55500 low RCS target mounting systems. It contains product descriptions, specifications and ordering information covering pylons and rotators. This system minimizes electromagnetic backscatter while providing a method for mounting and positioning RCS targets. A glossary of terms can also be found in the brochure. **Scientific-Atlanta, Inc., Atlanta, GA. INFO/CARD #218.**

Paper Describes Display Shielding

This technical paper, "A Study of Materials and Methods for EMI Shielding of Displays," was presented at the annual conference of the Society of Information Display. It provides a technical commentary and graphs with design considerations and constraints in the selection of EMI shielding windows for government, military, medical and sensitive business applications. **Teknit, Cranford, NJ. INFO/CARD #217.**

Test and Instrument Product Guide

United States Instrument Rentals has published its 1989/1990 product guide. It contains information on over 5,000 models from more than 170 manufacturers of electronic test and measurement instruments, data processing equipment and telecommunications test devices, available for rent, lease or sale. Included are specifications, descriptions, photos and other technical data. **United States Instrument Rentals, Inc., San Mateo, CA. INFO/CARD #215.**

Scalar Network Analyzer Brochure

This brochure describes the 562 scalar network analyzer that measures

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transmission, return loss and power over the 10 MHz to 40 GHz range. In addition to specifications, the brochure contains front panel functions, applications, accuracy tables and a list of available accessories. **Wiltron, Morgan Hill, CA. INFO/CARD #214.**

Application Note on Determining Bandwidth

Using Computer Simulation to Determine Operational Bandwidth of Digital Packages outlines the methods used to simulate the performance of a 196-pin VLSI package with commercially available software. A package representation that includes discontinuities is simulated using Touchstone[R]. Also, since the package is a passive device, the original model can be simplified and analyzed in the time domain. **EEsof, Inc., Westlake Village, CA. INFO/CARD #213.**

SAW Brochure

This brochure from Sawtek describes resonator theory of operation, one-port resonators, two-port resonators, delay lines, performance characteristics and packaging. A table lists performance parameters of the above products and a section on definitions is featured. **Sawtek, Inc., Orlando, FL. Please circle INFO/CARD #212.**

Product Catalog

This catalog, from Racal-Dana, covers programmable switching systems from DC to 26.5 GHz, universal counters/timers, digital test subsystems, and RF instrumentation. Specifications, photographs and ordering information is included. **Racal-Dana Instruments, Inc., Irvine, CA. INFO/CARD #211.**

SPICE Reference Guide

Simulating With SPICE features a tutorial on SPICE which allows the novice to become familiar with the program. Example problems, application note sections, benchmark circuits, a SPICE related bibliography and netlists of all circuits used are included. The book costs $64 and comes with a $50 IsSPICE coupon. For more information circle the reader service number. **Intusoft, San Pedro, CA. INFO/CARD #210.**

Contacts and Connectors Catalog

This catalog covers the AMP multiple COAXICON coaxial connector family, which incorporates a variety of male and female contacts into plastic rectangular housings or inserts with metal shells. It provides product facts, materials and finish information, performance specifications, and wire and cable selection charts. Information on application tooling, technical documents and connector hardware, as well as detailed product drawings, are included. **AMP Inc., Harrisburg, PA. INFO/CARD #209.**

Modulation Meter Bulletin

This six-page bulletin discusses the features of the Marconi Model 2305 50 kHz to 2.23 GHz modulation meter. Four block diagrams detail the operation, set-up, demodulation sequential signalling tones, and the measurement of audio tones. GPIB programming and programming codes are highlighted together with a specification summary on the instrument. **Marconi Instruments, Allendale, NJ. INFO/CARD #208.**

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1. Lotus spreadsheets from "Design of Constant Phase Difference Networks," by Thomas Keely.
2. Interpreter and compiled versions of "A BASIC Program for 90-Degree Allpass Networks," by Allan Lloyd.
3. HP BASIC program from "Bridged-Tee Delay Equalizers — A Computer-Aided Realization," by Robert Kane. (We are attempting to get this lengthy program translated into BASICA or GWBASIC.)
4. "Match," by Peter Martin, referenced in his article, "Design of Line Matching Networks," in the February 1989 issue.

Disk RFD-0389 (March 1989 RF Design)

1. "A Design Program for Butterworth Lowpass Filters"
2. "A Parallel-Coupled Resonator Filter Program"

Disk RFD-0289 (February 1989 RF Design)

1. "CAD for Lumped-Element Matching Circuits"
2. "Modeling PLL Tracking of Noisy Signals"

Prices are $9.00 each for a 5 1/4 in. diskette, or $10.00 for a 3 1/2 in. mini-floppy. Outside of the U.S. and Canada, add $8.00 (disks will be sent airmail). Make check or money order payable to *RF Design Software Service*. Specify disks wanted, and send orders to:

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