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ideas for engineers

November 1986

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Show Issue: New Products at RF Expo East

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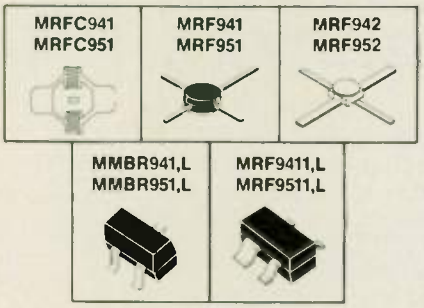

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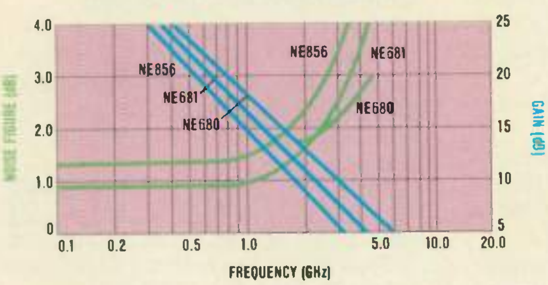

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November 1986

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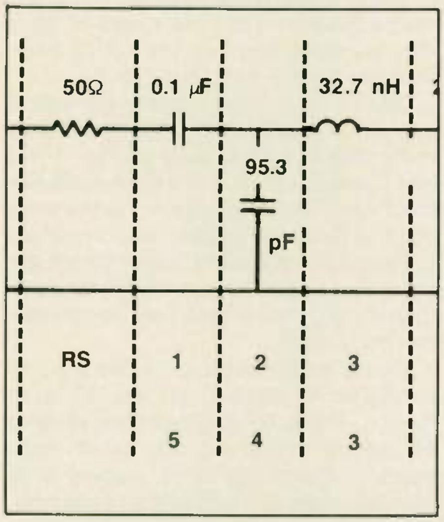

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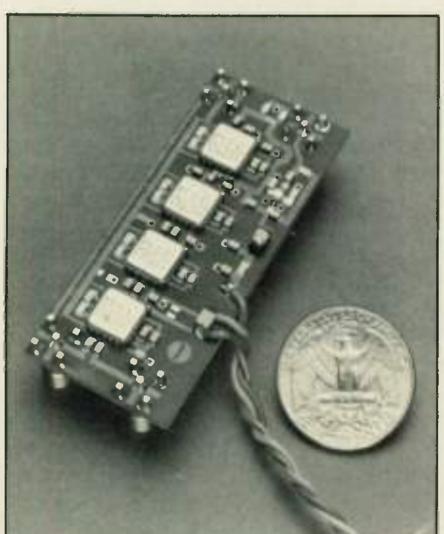

Page 84 — Log Amplifier Design

On The Cover

New Directions for Hewlett-Packard

In a significant change of direction, HP has entered the low-cost spectrum analyzer market, but with traditional HP quality. The HP 8590A covers 10 kHz to 1.5 GHz, weighs 30 lbs., and costs $9,500. More on pages 40-41.

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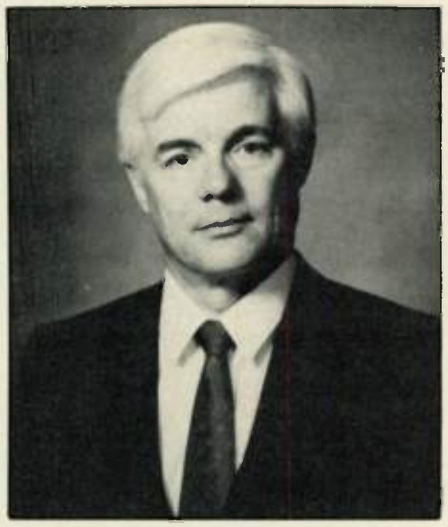
rf editorial

Thanks to the Contributors



By James N. MacDonald
Editor

Last month we used this space to take a bow and tell you about some of the ways we are proud to serve this industry. Like a conductor we now ask the "orchestra" to stand and be recognized. Without the writers who take the time to share their knowledge there would be no *RF Design*.

RF design engineering is one of the most demanding professions. The knowledge and mental agility required is such that many people could not master it, even if they wanted to devote the long hours of study. It is seldom an 8 to 5 job, and many RF engineers spend their free time doing some variation of what they do at work. We should all be grateful that some of them spend that time writing.

When we suggest that an engineer write an article about something he or she has developed, we sometimes hear the response, "Where would I find the time?" It is hard to answer that question. We know there probably is no extra time for writing articles. So, we have that much more appreciation for those who take the time from something else.

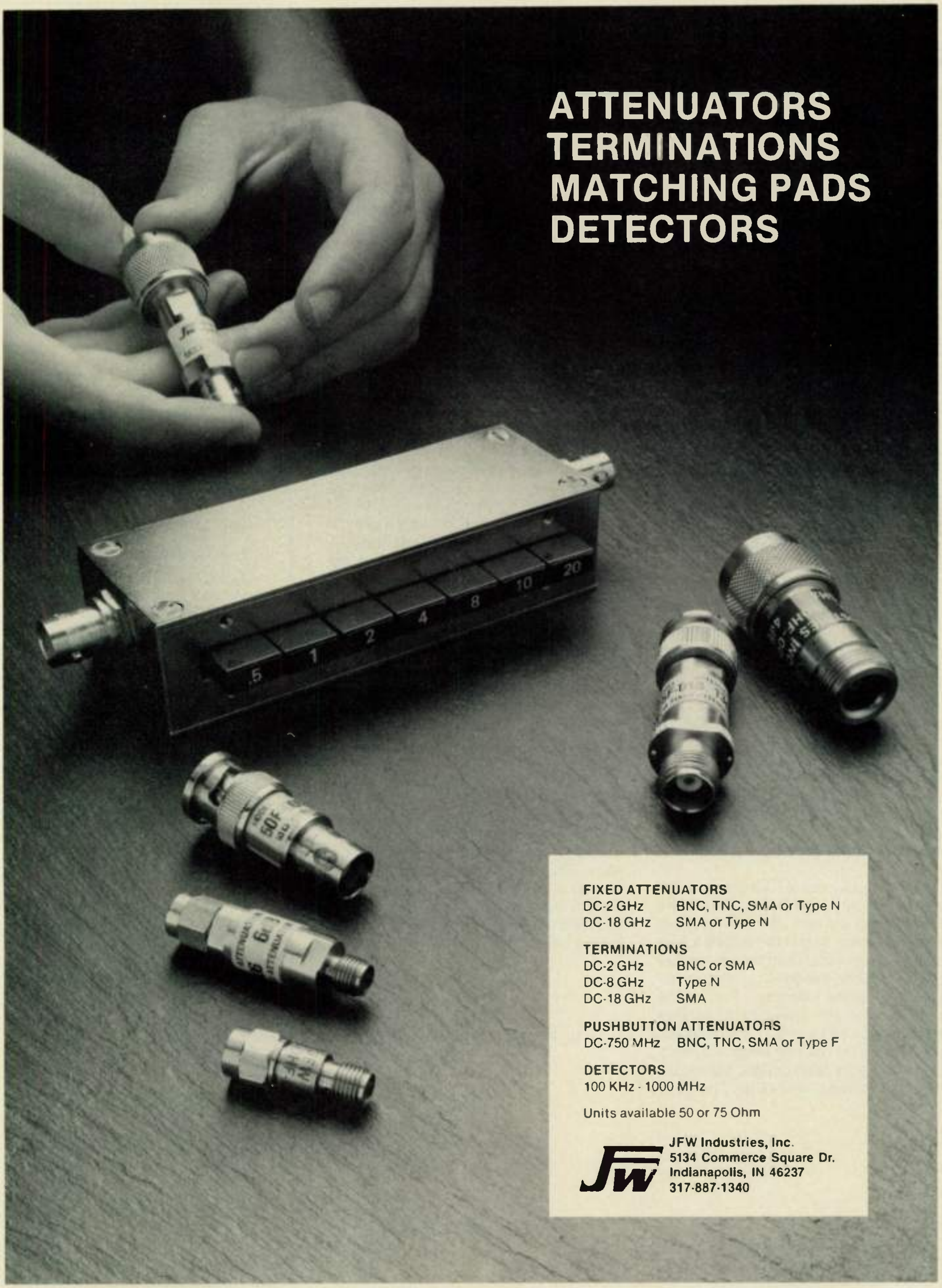
OK, it isn't all altruistic. We know our contributors like the recognition from readers and from their co-workers and supervisors. Sometimes they are asked to write an article to explain a new product. Behind it all, however, has to be the desire to share knowledge and advance the profession. They are proud of what they are doing and they want other engineers to be the best they can be.

One of the most gratifying aspects of this job is reading the Comment Cards that come in almost daily saying, "That was a terrific article. Let's have more like that." Or, "That article gave me the solution to a problem I've been working on for months." From talking with authors we know they hear such comments far more than we do. That is their real compensation for writing.

As this issue goes to press we are preparing for RF Expo East, Nov. 10-12, in Boston. There, RF engineers will present 58 papers on some aspect of RF engineering. Preparing these papers is at least as much work as writing an article, but this is the third show we have done in two years with such a technical program. The fourth will be next February, and the program is almost complete for that one. That will mean more than 200 different papers presented in a two year period. We sometimes wondered if we could keep the technical program that size, but the response to our call for papers gets larger with each show. And the quality of the papers remains high.

So, let's hear it for the people who write the articles and those who prepare the papers. Without them there would be no magazine and no show.

James N. MacDonald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The Power and the Glory

By Gary A. Breed
Technical Editor

1987 is going to be a fascinating year for RF engineers and *RF Design!* January's Special Report is on the subject of *RF in Scientific Research,* an area where RF engineering shares in the glory of radio astronomy, weather research, elementary particle physics, and space exploration. It has taken almost no effort to find many possible avenues to explore in preparing the report. RF is everywhere in research, from the expected voice and data communications to innovative measurements, analysis, and signal processing techniques where RF is integral to the project.

After the glory comes the power — RF power. February's *RF Design* will feature new techniques and new devices that make RF power one of the hottest topics in all of electronics, searching for higher power, broader bandwidths, higher frequencies, better efficiency. Part of the force pushing RF designers to achieve better power amplifiers is the research noted above, in Nuclear Magnetic Resonance research, laser drivers, particle accelerator "pumps" and radar. The other big push comes from communications systems, where reliability, efficiency and broad bandwidth are major goals.

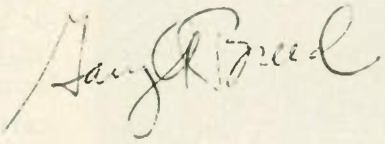
On a More Serious Note

Being attuned to scientific research for the January report, I noted an item that was carried in the papers a few weeks ago concerning the resignation of a brilliant young physicist at Lawrence Livermore Laboratories.

News about this country's top scientists is not unusual, but this story made me stop and think. Among the expected reasons for dissatisfaction (management style, job assignments) this man had another reason for leaving. He had brought his talents in the field of X-Ray lasers to the Laboratories with the expectation of using them in medical applications. Instead, he found himself working on "star wars" weapons, and chose not to continue that work.

I wondered, how many engineers stop and think about the consequences of their actions? Do they understand and accept the role of the finished product, beyond the assemblage of parts that is their piece of the system? Can they take pride when they see that it is useful, and do they become concerned when the end use doesn't fit their ideals?

Our recent survey indicated the median age of RF engineers to be about 36. I am 37 myself, and I can clearly remember the debate and question over every aspect of life during the tumult of the late '60s and early '70s. Have many of us kept even a little bit of that questioning attitude? Can we look beyond the power and the glory, and think about the responsibility we have for the use of our talents?



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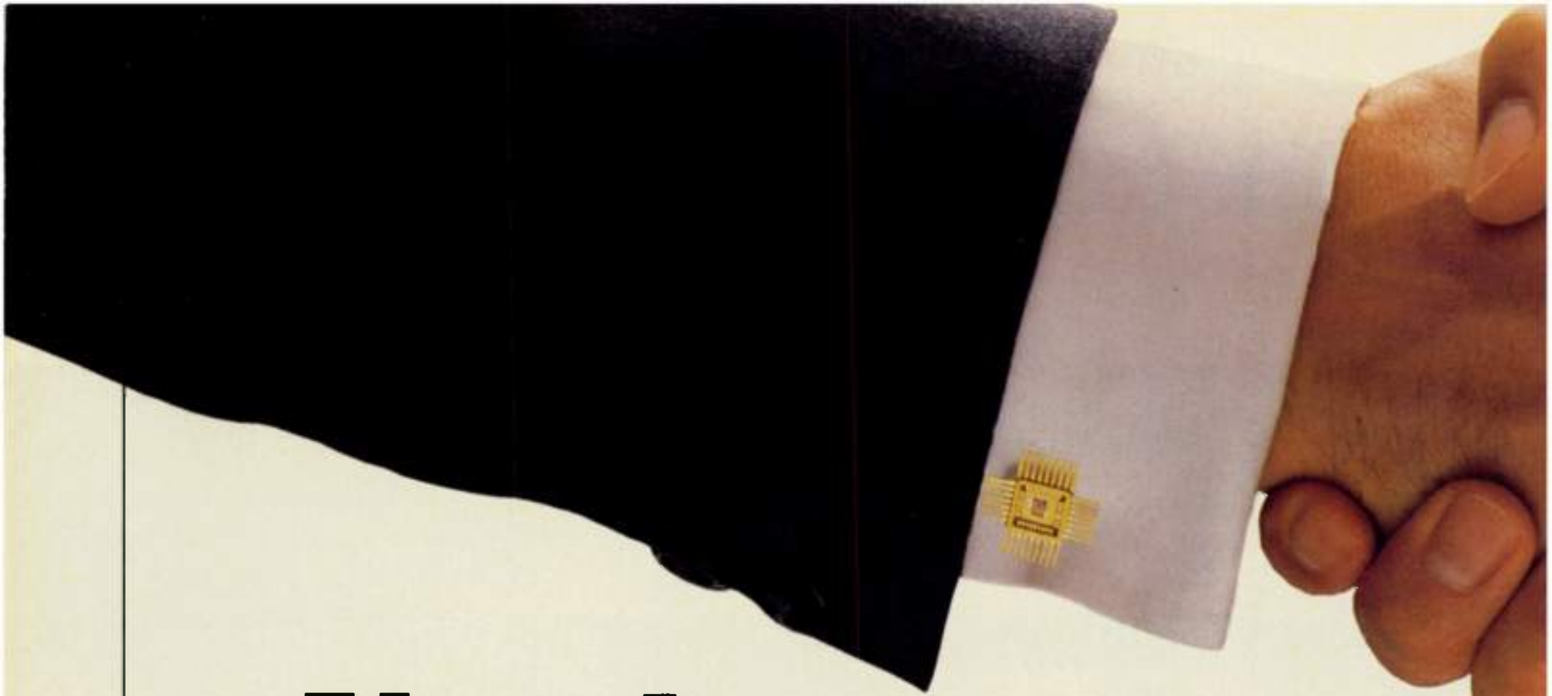

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rf letters

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Reed Relay Loss is Skin Effect

Editor:

Author Colin Gyles ("Anomolous Behavior of Reed Relays," September 1986, p.49) discusses a phenomena of hermetically sealed relays, excessive loss of high frequency response. Gyles correctly notes that relays employing magnetic materials exhibit far more loss than similar structures using non-magnetic materials.

His explanation of a magnetostrictive phenomenon stretches the imagination. His own photos show that the conductor's effective high frequency resistance is at least five times that of the DC value. Furthermore, the conductor seems to start out in the constricted or high resistance state before any current flows. It is hard to believe that the conductor constricts to one fifth of its final diameter in zero time.

The correct explanation is skin-effect loss. The magnitude of the loss is so much higher than we are used to seeing precisely because the materials used are magnetic. At high frequencies the current is pushed to the conductor's extreme outer surface by its high permeability.

Magnetic materials such as Kovar are used in the manufacture of hermetically sealed relays because they match the temperature coefficient of expansion of the glass to metal seals. Hence, both glass reed relays and TO-5 style metal relays exhibit the phenomenon — some more than others.

There are circuits which compensate for the phenomenon correctly, although good compensation requires more than a few components because skin effect is not a simple phenomenon which can be modeled accurately with a single pole-zero pair.

John Addis
Tektronix, Inc.
Beaverton, Ore.

Editor:

The article "Anomalous Behavior of Reed Relays" by Colin Gyles brings to your readers' attention an important phenomenon in precision wideband circuits. Mr. Gyles comments that the frequency variation of series impedance of these relays "can probably be attributed to magnetostriction." However, the phenomenon is most certainly due to skin effect.

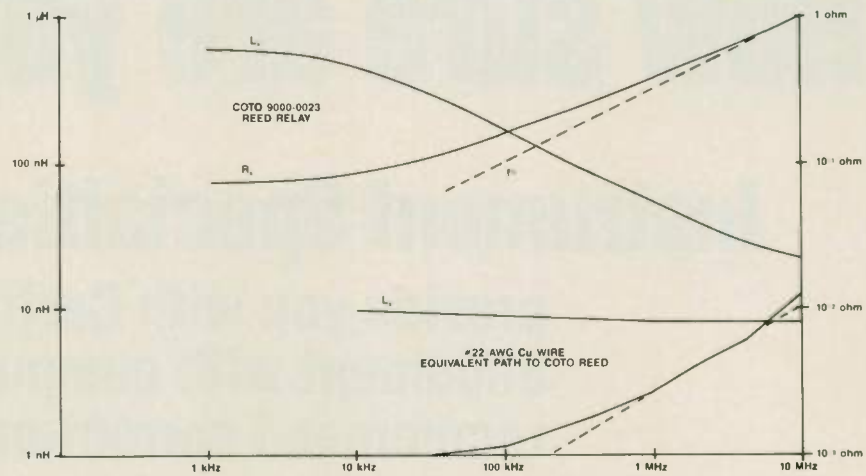

Figure 1.

Skin effect (1) in copper conductors is usually not of interest at low MHz frequencies unless the circuit impedances are very low and/or desired Q very high. The reed relay armature is made of a ferromagnetic material for which (mu/rho) is much higher than copper (mu = relative permeability, rho = resistivity). This lowers the frequency at which skin effect becomes dominant.

Measurements of Ls and Rs vs. frequency (f) are shown in Figure 1 for a COTO 9000-0023 reed relay and an equivalent length of #22 AWG Cu wire. These measurements were made directly on Hewlett-Packard 4375A and 4274A LCR meters, carefully autozeroed with a < 3 mm long copper short-circuit. At 4 MHz the skin effect in the reed is pronounced, causing most of the current to flow near the surface of the conductor. This increases Rs by almost an order of magnitude relative to 1 kHz. The graph clearly shows that Rs follows the SQRT(f) behavior associated with skin effect. Ls of the reed is considerably reduced at higher frequencies because the magnetic field is now mostly outside the high mu material. This is predicted by comparing the inductance of a solid cylinder with that of a cylindrical shell of the same outside diameter (2).

The elastic deformation implied by magnetostriction would require the length of a cylindrical reed to increase by a factor of 2.6 (assuming the volume of the reed remains constant) to obtain an increase of resistance by a factor of 7. Clearly this does not happen!

The use of ferromagnetic conductors in low impedance paths of precision wideband circuits should be avoided.

1. Reference Data For Radio Engineers, Sixth Edition (Howard W. Sams & Co., 1975), p. 6-4, 6-8.

2. L.J. Giacoletto, ed., Electronic Designers' Handbook, Second Edition, (McGraw-Hill, 1977), p. 3-43 item 4.

James J. Sanders, III
Teradyne, Inc.
Boston, Mass.

The Author's Response

Editor:

Thank you for a really nice presentation of my article on the "Anomolous Behavior of Reed Relays."

I have received a copy of John Addis's letter that he sent you, proposing that the reason for the observed anomoly is due to skin effect.

I have also had a call from Gordon Long of Tektronix who explained that they have been aware of this phenomenon for many years, attributing it to skin effect being magnified by the material's high permeability.

Since writing the article, I have documented similar effect in coaxial cables which are well documented to be attributed to skin effect.

Thus, although there might be a contribution caused by magnetostrictive effects, the consensus of opinion is that the predominant factor is skin effect.

Although this effect is well known at Tektronix, it is not generally known — indeed, the reed relay manufacturers don't mention it and two manufacturers contacted don't even know of the problem.

My reason for writing the article was to advertise this subtle problem to other engineers that might have trouble with it and not to prove the reason behind it. I state in the article that "the cause of this anomoly is *probably* due to magnetostriction;" no research or time was spent to verify it. In fact, another reason for publishing was to find out if anybody out there knew the reason, and to that end it was successful.

Colin Gyles
Analogic Data Precision Division
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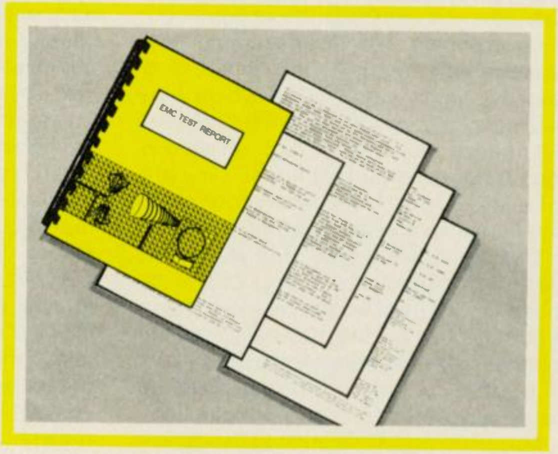

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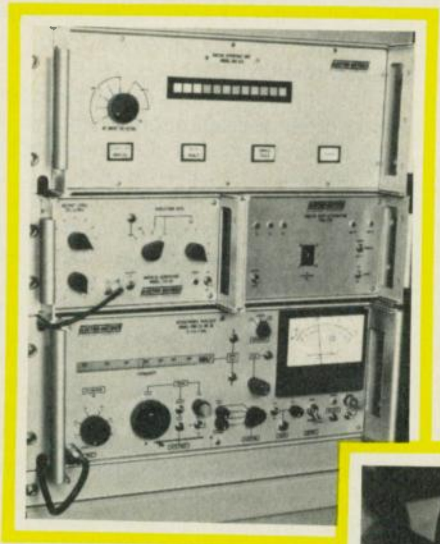
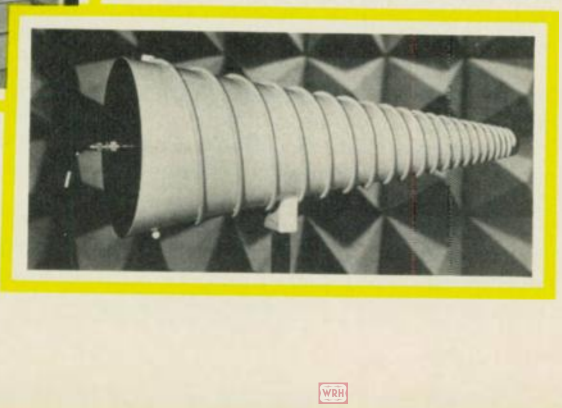


Tests are computer-controlled, to provide customized testing with precise results, repeatability, and hard-copy documentation to be included in your test reports. In addition to those tests mentioned above, shielding can be tested to MIL-STD-285, IEEE 188, NSA, etc. RF gaskets are tested using the SAE ARP-1705 transfer impedence method, to 140 dB. Computers and proved software, plotters, signal sources, oscilloscopes, meters, receivers, amplifiers and antennas are employed in a quantity and degree of sophistication not available even to most manufacturers of Class A and B devices. In fact, few, if any, other test facilities in the U.S. have a comparable array of equipment, calibrated and traceable to the National Bureau of Standards!



rf letters *Continued*

Well Paid Compared to Whom?

Editor:

As I read your article "Profile of the RF Engineer," some questions came to mind. It is possible that those who respond to your survey do not represent a valid sample of RF engineers. The fact they receive your magazine would indicate they are keeping track of developments in their field, and therefore probably the elite of the group. I recall that back in 1936 the "Literary Digest," a leading business publication of its day, polled its readers regarding the result of the presidential election. Based on this poll they predicted Alf Landon would defeat FDR. As you may remember, Landon won only Maine and Vermont. Also, it appears that the RF engineers employed by the federal government (the largest employer of engineers) were not adequately represented. Starting salaries are below $25,000 and only a small portion of government engineers ever earn as much as $45,000/yr.

When you state RF engineers are well paid, I am reminded of the question "How is your wife?" and the standard answer, "Compared to who?" RF engineers are not well paid when compared to those in the legal, medical, or business professions. They are not even well paid when compared to public school teachers. Recent figures show that the average public school teacher in California receives $29,950 per year for teaching less than 1200 hours. This equates to $50,000/yr if they worked as many hours as engineers.

While engineers do earn a living wage, consider the caliber of these engineers.

1. Scholastically they were in the top 10 percent of their high school class.
2. They completed perhaps the most difficult and time consuming of all degrees.
3. They spent four years and $50,000 learning their trade. (No, they did not receive an education.) This places them some $100,000 behind union craftsmen who elected to learn their trade on the job.
4. They struggle constantly trying to keep abreast of developments in engineering. For example, when I graduated in 1949 there were no semiconductors, satellite communications, digital computers, color television, information theory, etc. The older you become, the more difficult the task of keeping up becomes. It is like running the 220-yard dash. You run as fast as you can for the first 200 yards, and then you sprint.
5. They have the misfortune of living in the only country in the world where engineers are paid essentially the same wage as steelworkers, the building trades, etc.

The fact that RF engineers are relatively young can be accounted for in two ways. First, once they are past 40, they are vulnerable to replacement due to obsolescence. Second, many of the brightest ones leave the profession in search of greener pastures.

Carl W. Chapman
Carmichael, Calif.

Surveys Show Changes

Editor:

It was interesting to compare your recent survey results with those in 1982.

	1982	1986
Median Income	32K	45K
Engineering Exp.	15 yrs.	12 yrs.
Education:		
BSEE	40%	66%
POST GRAD	29%	37%
BS (any)	32%	17%
NO DEGREE	9%	10%
Amateur Radio License	40%	33%

Median income for the four-year period has seen the expected increase.

Experience levels have dropped slightly from 15 to 12 years in engineering with a much bigger drop in RF engineering experience from 11 to six years.

This seems to imply entry of experienced engineers into the RF field rather than an increase in young engineers entering the field.

There seems to be a dramatic increase in BSEEs vs. BS degree holders though survey techniques might account for some of this.

Non-degreed RF engineers still remain around the 10 percent level.

The amateur radio percentage has dropped from 40 percent to 33 percent, possibly also indicating entry of engineers into RF from other disciplines.

The 33 percent figure is actually much *higher* than his "ham" operator figured. Your suggestion that amateur radio is not as big a factor as one might think is somewhat like saying that California's 15 percent of the total U.S. population is not much of a factor in U.S. government or culture!

It is also interesting to note the strong correlation between the MIL/Aerospace percentages and those who have access to IBM PCs or "stand-alone workstations" (73 percent and 72 percent, respectively). It would be interesting to look at the same figures for industrial and consumer electronics segments to see just where those "other" computers are being used.

It is also interesting to note that almost 20 percent either "share" or have "none" . . . one out of five *RF Design* readers!

James Eagleson
RF Engineer
Identronix, Inc.
Watsonville, Calif.

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1681-35	1600-1800 MHz	35 W	7.0 dB	28 Volts
LDR55LA	1200-1400 MHz	55 W	6.6 dB	36 Volts
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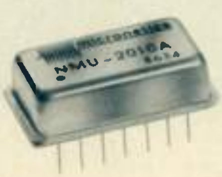
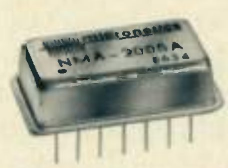


- Package:
  14 Pin DIP
- Output:
  5dB, 15dB, 30dB ENR
- Upper Freq. Limit:
  500MHz to 1.5GHz
- Pulse:
  TTL Input

INFO/CARD 105

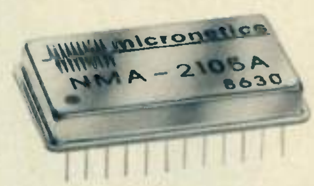
μ-NOISE
NMA 2000 Series



- Package:
  14 Pin DIP
- Output:
  10 millivolts RMS
- Upper Freq. Limit:
  100KHz to 500MHz

INFO/CARD 106

μ-NOISE
NMA 2100 Series



- Package:
  24 PIN DIP
- Output:
  1 volt RMS
- Upper Freq. Limit:
  100KHz to 500MHz

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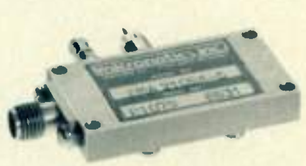
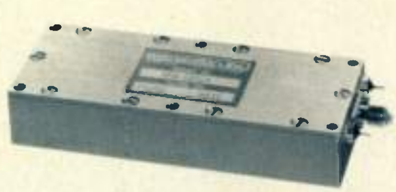


- Package:
  1.5 x .8 x .25 inches
- Output:
  1 volt RMS
- Upper Freq. Limit:
  100KHz to 500MHz

INFO/CARD 12

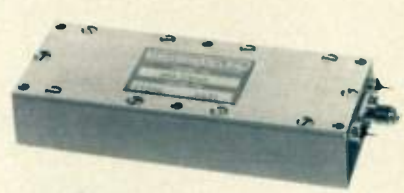
μ-NOISE NMA 2200 Series



- Package:
  3.75 x 1.5 x .7 inches
- Output:
  +13dBm
- Upper Freq. Limit:
  100KHz to 300MHz
- Output VSWR:
  1.2 to 1

INFO/CARD 103

μ-NOISE NMA 2300 Series



- Package:
  3.75 x 1.5 x .7 inches
- Output:
  3 Watts Peak
- Upper Freq. Limit:
  100KHz to 300MHz

INFO/CARD 104

rf calendar

November 10-12, 1986
RF Expo East
Boston Marriott Copley Square, Boston, Massachusetts
Information: Kathy Kriner, Convention Manager, Cardiff Publishing Co., 6530 So. Yosemite St., Englewood, CO 80111; Tel: (303) 694-1522 or (800) 525-9154

November 12-13, 1986
FALLCON '86 — 35th Annual IEEE Fall Conference Electronics Exposition
Five Seasons Center, Cedar Rapids, Iowa
Information: Barry F. Lipp, Publicity Coordinator, P.O. Box 451, Marion, IA 52302; Tel: (319) 395-8545

November 18-21, 1986
Wescon/86 High Technology Electronics Exhibition and Convention
Anaheim Convention Center, Anaheim, California
Information: J. Fossler, Electronic Conventions Management, 8110 Airport Blvd., Los Angeles, CA 90045; Tel: (312) 299-9311

December 4-5, 1986
28th Automatic RF Techniques Group Conference
Don CeSar Beach Resort, St. Petersburg, Florida
Information: Richard Irwin, Systems for Automatic Test, 1282 Reamwood, Sunnyvale, CA 94089; Tel: (408) 734-9447

January 12-15, 1987
SMART III
Hyatt Regency, New Orleans, Louisiana
Information: Electronics Industries Association, 2001 Eye St. N.W., Washington, D.C. 20006; Tel. (317) 261-1306

February 11-13, 1987
RF Technology Expo 87
Disneyland Hotel, Anaheim, California
Information: Kathy Kriner, Convention Manager, Cardiff Publishing Co. (see address above)

February 25-27, 1987
Industry-University Advanced Materials Conference
Colorado School of Mines, Golden, Colorado
Information: Dr. Jerome G. Morse, Advanced Materials Institute, Colorado School of Mines, Golden, CO 80401; Tel: (303) 273-3852

April 1-8, 1987
Electronics and Electrical Engineering '87
Hannover Fairgrounds, Hannover, West Germany
Information: Hannover Fairs USA, Inc., 103 Carnegie Center, P.O. Box 7066, Princeton, NJ 08540; Tel: (609) 987-1202

April 21-23, 1987
Electrical Overstress Exposition
San Jose Convention Center, San Jose, California
Information: Jim Russell, EOE, 2504 N. Tamiami Trail, Nokomis, FL 33555; Tel: (813) 966-9521

May 11-13, 1987
37th Electronics Components Conference
Boston Park Plaza Hotel and Towers, Boston, Massachusetts
Information: Tom Pilcher, Electronic Industries Association (see address above)

rf courses

The George Washington University

Radar Systems and Technology
November 3-7, 1986, Washington, DC

Grounding, Bonding, Shielding and Transient Protection
December 8-11, 1986, Ottawa, Canada

Mobile Cellular Telecommunications Systems
December 8-10, 1986, Washington, DC

Wideband Communications Systems
December 15-19, 1986, Washington, DC

SAW Devices and Their Signal Processing Applications
February 23-26, 1987, Washington, DC

Frequency Synthesis
February 9-11, 1987, Washington, DC

Frequency Hopping Signals and Systems
February 23-25, 1987, Washington, DC

Spread Spectrum Communication Systems
March 9-13, 1987, Washington, DC

Synchronization in Spread Spectrum Systems
April 6-10, 1987, Washington, DC

Information: Merril Ann Ferber, Assistant Director, Continuing Education Engineering Program, The George Washington University, Washington, DC 20052; Tel: (800) 424-9773

Georgia Institute of Technology

Principles of Modern Radar
November 3-7, 1986, Atlanta, Georgia

Phased-Array Antennas: Theory, Design and Technology
November 18-21, 1986, Atlanta, Georgia

Information: Diedre Mercer, Department of Continuing Education, Georgia Institute of Technology, Atlanta, GA 30332; Tel: (404) 894-2547

Besser Associates, Inc.
Principles of RF and Microwave Circuit Design
November 12-14, 1986, Baltimore, Maryland
December 15-17, 1986, Santa Clara, California

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

Interference Control Technologies, Inc.
TEMPEST Design, Control and Testing
November 4-7, 1986, Santa Rosa, California

Grounding and Shielding
November 11-14, 1986, San Diego, California
December 2-5, 1986, Washington, D.C.

TEMPEST Facilities Design, Installation and Operation
November 11-14, 1986, Washington, D.C.
January 27-30, 1987, San Jose, California

Practical EMI Fixes
December 9-12, 1986, Orlando, Florida
January 20-23, 1987, San Diego, California

Information: ICT, P.O. Box D, Gainsville, VA 22065

Plessey Technology to Power Wristwatch Communications Terminal

A wristwatch that can receive and display broadcast messages has been developed by AT&E Laboratories, Beaverton, Ore., a subsidiary of AT&E Corporation, San Francisco. The wristwatch will use advanced integrated circuit and radio technology from UK-headquartered Plessey Semiconductors.

AT&E Laboratories selected Plessey to integrate AT&E proprietary system technology into two proprietary semiconductor chips that will be used in an average-size wristwatch, to receive and decode a radio signal and display a digital message, such as "Call Office," or call any designated phone number, as well as keep time.

The personal wireless wristwatch communications terminal, known as the Receptor™, combines an extremely sophisticated FM radio receiver and a highly accurate standard time piece. Not only are messages broadcast by local FM radio stations received and displayed anywhere in the world, but also variations in local time, such as those caused by travel between time zones, are automatically corrected for using coded data on the FM subchannel signal.

"Plessey has the system knowledge, expertise in RF communications, and the capability to implement those systems into integrated circuits, as well as high speed, low power bipolar and CMOS processes that will allow the devices to be economically produced for operation in a wristwatch/radio environment," said Joseph F. Stiley, III, president of AT&E Labs.

Plessey will supply two proprietary circuits, a bipolar direct conversion receiver and a CMOS decoder for extracting the message signal for further processing and display. The RF receiver must operate with a very low power input (−92 dB) from the watchband antenna, have excellent dynamic range (40 to 60 dB) and operate at low power and voltage. The CMOS device handles the interface between the receiver and a microprocessor which processes and displays the message to the user.

IBM Ranks No. 1 in Europe

According to a recently published analysis from Benn Electronics Publications Ltd. (BEP), IBM has retained its position as the leading electronics company operating in the European market. As monitored in the latest Seventh Edition of Benn's European Electronics Companies FILE, Top 100 Table, IBM achieved electronic product sales in Europe of over $11.7 billion in the 1984/85 period, compared to $11.0 billion the previous year.

This is the second consecutive year that Philips, the largest European-owned electronics group, has been forced into second place by the U.S. computer conglomerate. For several years, Philips exerted its leadership in the marketplace and for the period 1980/81 to 1982/83 held on confidently to the leading position with sales of over $10 billion. However, as a result of a depreciation in the guilder against the dollar, Philips' electronics sales have fallen below the $10 billion mark in the 1983/84 periods, to $9.7 billion and $9.8 billion, respectively, thus making way for the challenge from IBM.

Seventy percent of the top 20 electronics companies operating in Europe are European-owned, and unlike Philips most of them have improved on their overall ranking when compared with the previous year. Seven of the 14 companies moved up the rankings, five moved down and two stayed the same. CGE of France and STC of the U.K. made marked improvements in their ranking by acquisition, rising 10 and 16 positions, respectively. CGE rose to No. 7 as a result of the restructuring of the French electronics industry, which led to CGE acquiring the communications operations of Thomson. STC rose to No. 10 in the rankings after its acquisition of ICL, the £1 billion British computer group.

In the U.S.-owned group there was very little change with two groups, DEC and Hewlett-Packard, moving up the rankings, two groups moving down (ITT and Xerox) and IBM remaining unchanged at the top. The sole Japanese representative in the top 20 was Matsushita, which dropped two places to No. 12.

Overall, the total electronics-only sales of Europe's leading electronic groups, as monitored in the Seventh Edition of Benn's *European Electronics Companies File*, reached $104.6 billion in 1984/85, representing 30 percent of world electronics output (as monitored in BEP's *Analysis Series* database of 30 leading electronics nations).

The most profitable companies were either American or Japanese-owned companies, with average pre-tax profitability of these two groups reaching 16.9 percent of sales, and profit per employee averaging $16,340.

The only European-owned group to rival this dominance is Norsk Data of Norway with an average pre-tax profitability

Table 1. Top 20 electronics companies operating in Europe.

Rank 1984/85	Results to 1983/84	Month/Year	Company	Country	1984/85 Electronics Sales ($M)
1	1	12/84	IBM	US	11,741*
2	2	12/84	Philips	Netherlands	9,845
3	3	9/84	Siemens	W. Germany	8,407
4	5	12/84	Thomson	France	5,337
5	6	3/85	GEC	UK	4,013
6	4	12/84	ITT	US	3,576*
7	17	12/84	CGE	France	2,883
8	7	12/84	LM Ericsson	Sweden	2,854
9	8	12/84	Robert Bosch	W. Germany	2,745
10	26	12/84	STC	UK	2,145
11	13	6/85	DEC	US	1,978*
12	10	11/84	Matsushita	Japan	1,966*
13	14	10/84	Hewlett-Packard	US	1,765*
14	9	3/85	Plessey	UK	1,725
15	16	12/84	Olivetti	Italy	1,583
16	12	12/84	Xerox	US	1,582*
17	11	12/84	Groupe Bull	France	1,556
18	21	12/84	STET-IRI	Italy	1,396
19	20	12/84	Nixdorf	W. Germany	1,339
20	18	3/85	Thorn EMI	UK	1,336

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reaching 17 percent of sales. By comparison with the large electronic multinationals, Norsk Data is a relatively small company with overall electronics sales of $168 million, ranking it No. 86 in the Top 100 table. However, irrespective of size the group also managed to lead the European-owned group in terms of the highest profit per employee with a figure of $12,897, while no other European electronics company managed to return a profit of over $10,000 per employee.

Further information on this and other BEP publications is available from: Beryl Hay or Jackie Amos, Benn Electronics Publications Limited, Chiltern House, 146 Midland Road, Luton, Beds, LU2 OBL, England, Telephone (0582) 421981, Telex 827648 BENNLU G, or please circle INFO/CARD #200.

Magnavox Introduces Automatic Vehicle Location System

A newly developed automatic vehicle location system (AVLS), introduced by Magnavox Advanced Products & Systems Company, Torrance, Calif., provides an automated technique for locating, monitoring, communicating with and managing vehicle fleets.

Gerald A. Gutman, president of the Magnavox subsidiary NAV-COM Inc., said the AVLS can help emergency services improve response time and reduce demands on manpower, while keeping initial capital investment to a minimum. Gutman noted that this technology may also be suitable for a host of other applications, such as security vehicles, armored trucks, dangerous cargo transportation, public utilities, and even motor freight and package delivery services.

The Magnavox system uses compact navigation sensors and processors in each vehicle. These are linked via radio to a central computer at the dispatcher's station. There, the location and status of each vehicle are shown on a large color video map display.

A unique feature of the Magnavox AVLS is its use of precise automatic "dead-reckoning" techniques to track the vehicles' positions. The system does not depend on constant reception of signals from radionavigation sources, such as Loran-C, which may be subject to disruption, interference and error. Position sensors track the vehicle's movement by automatic inputs from a compass and speedometer. Position data is automatically recalibrated periodically with signals from navigational satellites.

The dispatcher's computer monitors each vehicle's location by sending out a short data burst triggering an automatic response from the vehicle's processor. Normally, the driver will not know when the vehicle is being polled. The frequency with which a given vehicle is polled is determined automatically by the dispatcher's computer and varies according to that vehicle's current status. Thus, a vehicle that is moving very fast is polled more often than one moving slowly. The polling interval changes automatically as the vehicle status changes. "This type of smart adaptive polling helps to optimize the radio link, minimizing congestion and keeping the frequency channel open for essential traffic," said Gutman.

Pre-programmed status messages can be included in position reports. These messages can be initiated by the driver at the touch of a button. Thus, the dispatcher can learn when the driver arrives at the scene, when the driver leaves his vehicle or if a driver is in trouble.

The system can also provide a wealth of information for planning and analyzing fleet activities. Vehicle locations, ID, time, status and other pertinent data can be stored in a database for future analysis.

For further information, contact Theresa Dale, Marketing Services Department, Magnavox Advanced Products and Systems Company, 2829 Maricopa Street, Torrance, Calif. 90503, telephone (213) 618-1200, telex 696101, or circle INFO/CARD #199.

EMC Labs Accredited by NBS

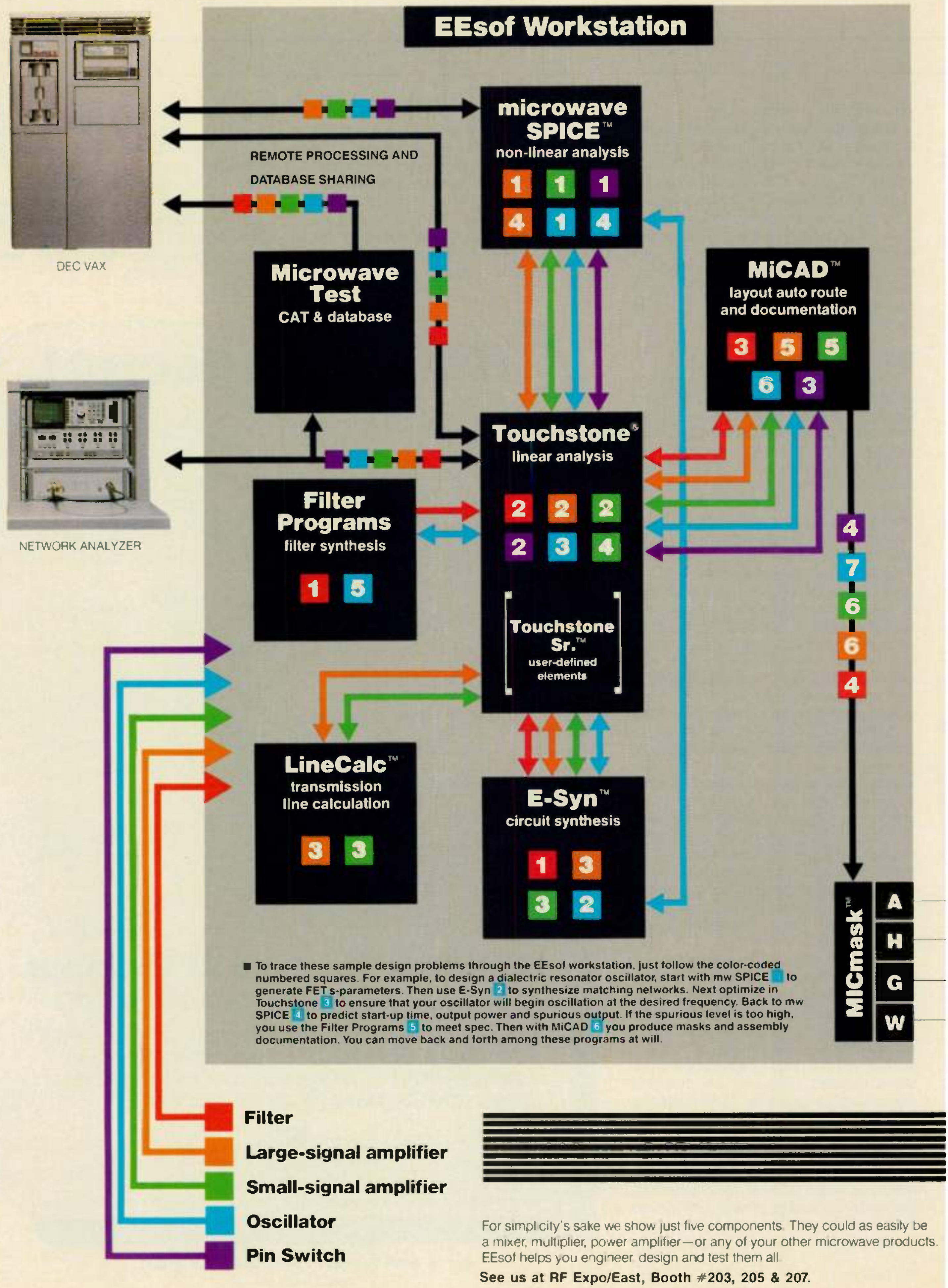
Fourteen private sector laboratories that perform electromagnetic compatibility (EMC) and telecommunications equipment testing have been accredited under the National Voluntary Laboratory Accreditation Program (NVLAP), established by the Commerce Department's National Bureau of Standards (NBS). The laboratories were accredited for selected test methods under a new electromagnetics laboratory accreditation program (LAP). This LAP was established by NBS — which manages NVLAP — at the request of five commercial testing laboratories seeking international recognition for EMC accreditation. International recognition of U.S. laboratories and test methods has been a high priority of industry groups and manufacturers to aid them in exporting their products to foreign countries.

With NVLAP accreditation, the laboratories automatically receive international recognition for their testing services through NBS' agreements with the United Kingdom's National Measurement Accreditation Service, Australia's National Association of Testing Authorities, and New Zealand's Testing Laboratory Registration Council. Under these agreements, test data reports listed by an accredited laboratory in one system are recognized by the other national accreditation systems.

Under NVLAP procedures, laboratories can apply for accreditation in one or more of the recognized test methods that make up the electromagnetics LAP. The LAP provides recognition to accredited laboratories that are capable of performing specific test methods for conducted emissions, radiated emissions, and terminal equipment compatibility in accordance with Federal Communications Commission (FCC) standards.

Established in 1976, NVLAP is a voluntary system whereby organizations and individuals request NBS to establish a laboratory accreditation program. On an individual basis, laboratories seek accreditation for having the competence to use specific test methods. "Competence" is determined by evaluating applicant laboratories to assure that they have the equipment, staff, and procedures necessary to perform recognized tests in accordance with nationally or internationally accepted standards or test methods. NVLAP-accredited laboratories pay annual fees, go through on-site reassessment every two years, and participate in scheduled proficiency testing to maintain accredited status. The laboratories are listed in the NVLAP directory that is distributed worldwide.

For further information or a list of accredited laboratories, contact: Manager, Laboratory Accreditation, A531 Administration Building, National Bureau of Standards, Gaithersburg, Md. 20899, telephone (301) 921-3431.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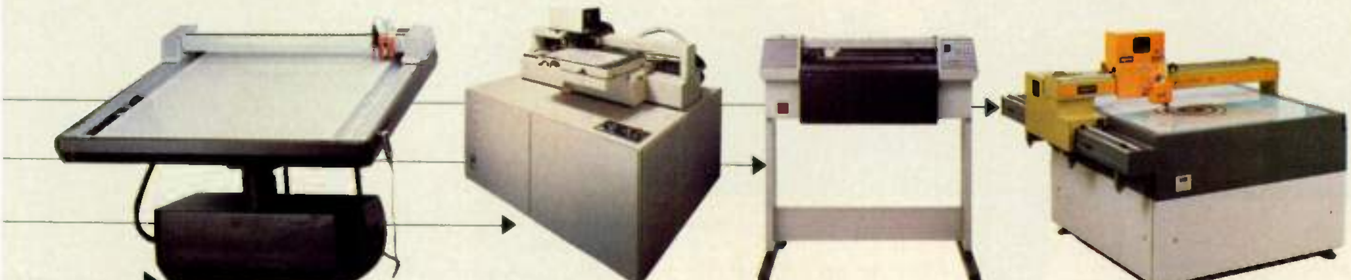

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New Shielding Methods and Materials

By James N. MacDonald
Editor

With the proliferation of radio communications equipment, shielding has become a serious problem. Manufacturers must protect their equipment from interference and must assure that it does not radiate excessive RFI. The more powerful equipment sometimes must be surrounded by shielding walls to contain its radiation and protect it from interference. Even entire buildings are shielded for security reasons. Signals radiated from computers, for example, can be detected from a surprising distance outside the building and read.

In this Special Report we take a short look at some of the new shielding materials and methods on the market to give RF engineers an idea of the possibilities available. Shielded enclosures and shielding materials have been discussed in this magazine before, so we will not attempt to cover everything on the market, only some new developments.

Much recent research is devoted to shielded rooms. Such rooms are important for measuring radiated emissions and testing protected equipment and for such medical techniques as magnetic resonance imaging. Lining the walls with conductive foil or metal is not difficult, but RF radiation can penetrate seemingly tight cracks. Keeping it in or out is not as easy as was once thought. Major construction problems arise when a prefabricated structure is assembled in a hospital, plant or elsewhere.

Sealing Doors

Joints and doors pose major difficulties. Chomerics, Inc., Woburn, Mass., recently announced METALKLIP, clip-on gasketing to seal doors. The clip-on feature is said to provide a simple means of mounting a carbon-loaded silicone or wire mesh EMI gasket strip on sheet metal enclosures. The corrosion-resistant stainless steel spring clip in which the gasket material is secured snaps onto the edge of a door or cabinet frame. Sharp teeth or tines inside the clip ensure electrical continuity between the METALKLIP assembly and the panel and bite through non-conductive paint.

METALKLIP gasketing can be fabricated with a broad range of integral EMI gasket materials, the company said, including carbon-loaded elastomer, Monel and Ferrex over a silicone tube core, or Monel mesh over a stainless steel mesh core. Solid or sponge neoprene and silicone cores are also available. It is offered in standard 8-foot lengths or custom cut. For additional information circle INFO/CARD #220.

Shielded Enclosures

For temporary shielding, a British company has announced a silver coated nylon mesh that can be an inexpensive screen to protect sensitive equipment and contain radiated emissions. Chemring Ltd., Portsmouth, England, said the material can be made into curtains by normal sewing techniques or bonded into fiberglass laminates for interior surfaces. For information circle INFO/CARD #195.

As this issue is distributed, Keene Corp., Ray Proof Division, Norwalk, Conn., is announcing a new 60 dB modular shielding system. Series 73 Plus is modeled after NSA requirements 73-2A with a greater shielding effectiveness. Bill McClain, vice president for sales, said the shielding industry has been hit with a series of new shielding requirements lately.

"When they start totaling up the area of shielding required and total number of dollars they start saying there's got to be a cheaper way to do it," McClain said.

He said a lot of customers have been opting for low performance, with such alternatives to welded or modular construction as foil backed sheet rock, screen and spray on.

"The problem is that they think they can get a drywall contractor to install the foil backed sheetrock. Customers have been seeing all sorts of problems."

McClain said the contractors are not guaranteeing shielding performance. He said he knows of many applications where it has not worked.

"It just has not been around long enough," he said. "In a laboratory under perfect conditions it works. Contractors are not used to the fact that if you scratch it, you've lost your shielding performance. They are not used to the care that has to be taken with penetrations. When you look at the specs for drywall construction it is horrendous. It is terribly labor intensive."

McClain described one installation where the contractor began with 12 people and now has 54 on the job.

Ray Proof installs the modular panels they have just announced. The Series 73 Plus is like the Series 81 they have been installing, except that the 28-gauge steel plate is only on one side. The structural strength comes from the framing and 3/4 inch Novaply.

"We get some fairly significant cost savings," he said. "We can guarantee not only the shielding effectiveness, but also the schedule. We know exactly how much it is going to cost and exactly how long it is going to take. Shielding does not become the critical path on the project."

Circle INFO/CARD #198 for additional information about Ray Proof shielded walls.

New Connector Shielding Method

G&G Technology Inc., Camarillo, Calif., has developed a method of terminating a shielded twisted pair into the back of a Series 4 type connector. Mike van Brunt, manager of the Electromagnetics Laboratory, said this method provides shielding all the way to the face of the grommet. It was developed for the U.S. Navy to eliminate the 7 inch pigtails now used.

"We solder a specially designed contact to the shield, and the contact plugs into a groundplane at the back of the connector," van Brunt said. "An important aspect of the design is that it is field repairable."

He said the company has received a patent and is going into the prototype and testing stage with the device. The company expects to market it within six months.

A New Mica Compound

Wilson-Fiberfil International, Evansville, Ind., has just introduced the industry's first nickel-coated mica compounds, which extend the performance spectrum of elec-

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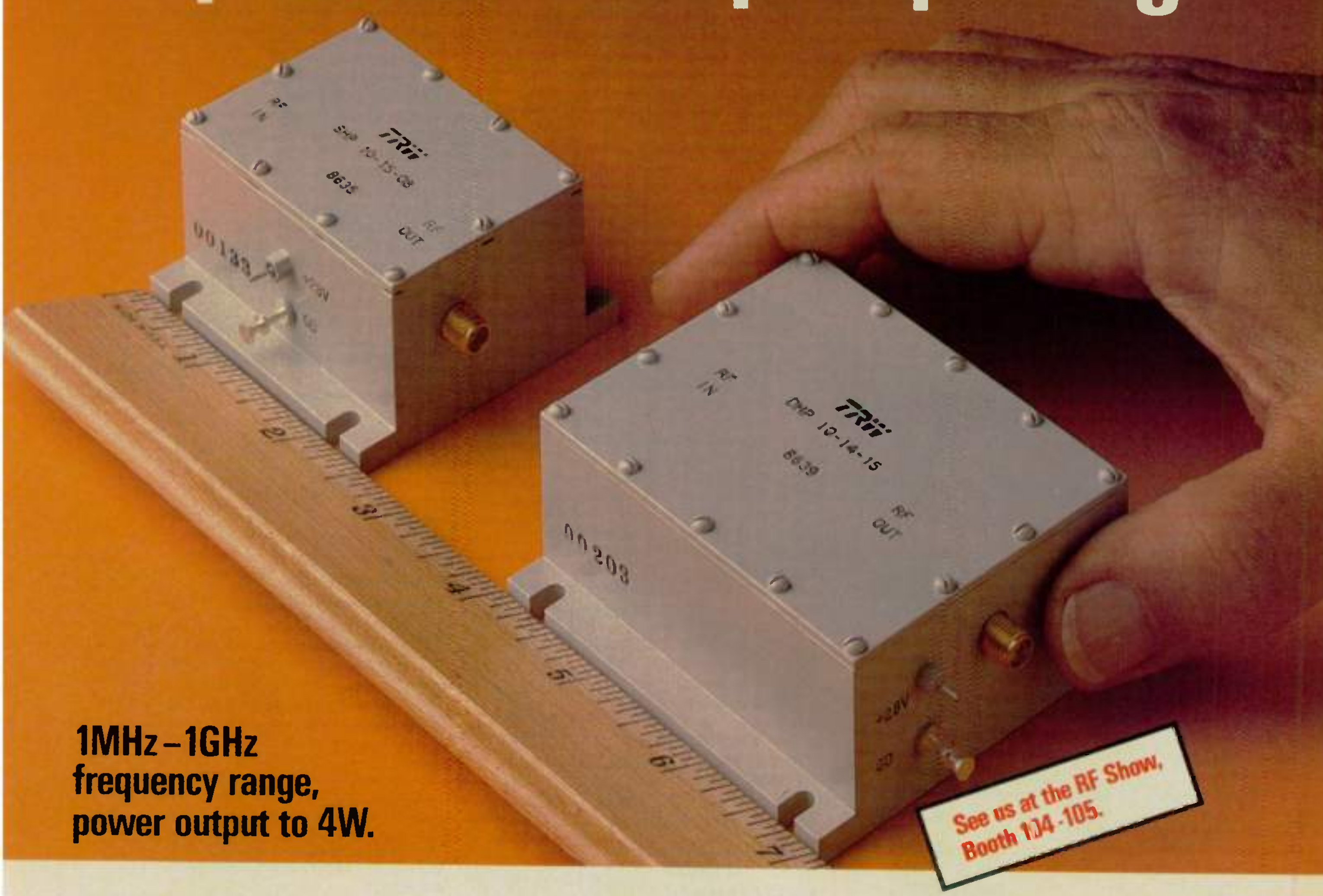

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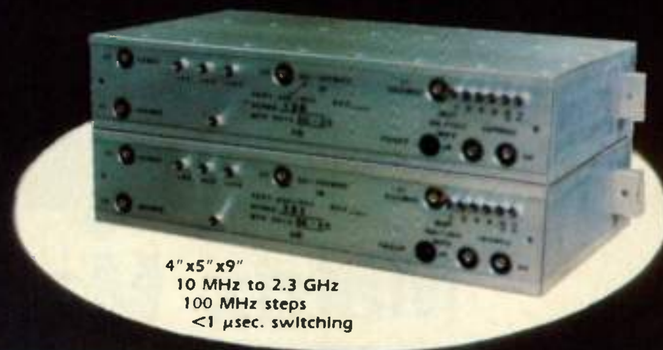

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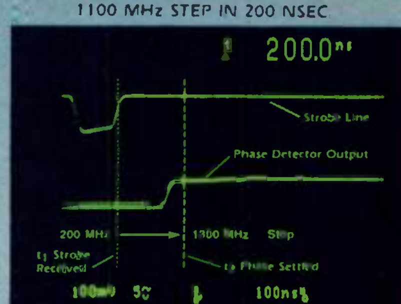

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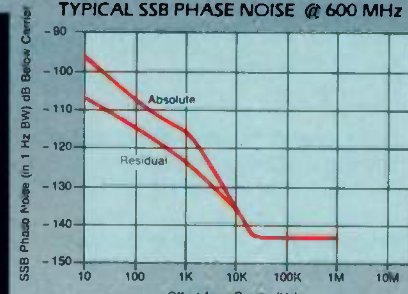

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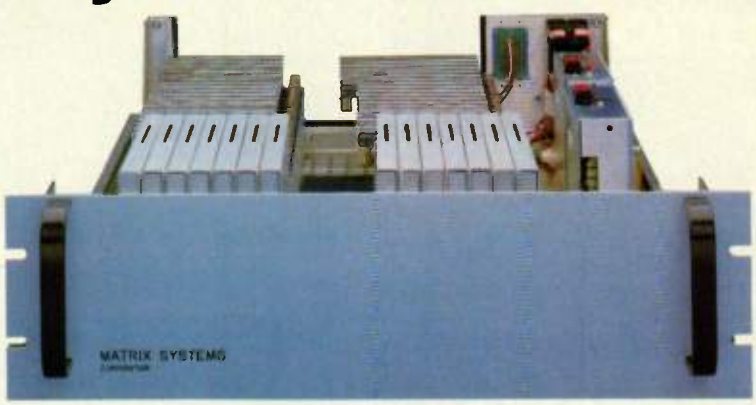

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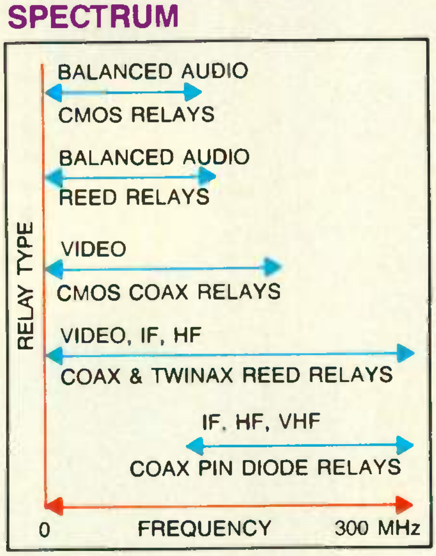

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DMOS Analog Switches

Solid-State RF Switching Seeks New Applications

By Gary A. Breed
Technical Editor

Analog switching for DC and low frequencies has become quite commonplace, with many standard CMOS components now available. However, CMOS does not have the low capacitance and ON resistance needed for effective RF switching. Despite these shortcomings, CMOS has found its way into many RF and video applications up to several MHz. Demand for the convenience of digital control and the reliability of solid state is evident.

DMOS (double-diffused MOS) technology has now reached a level of maturity for standard RF switching products to be readily available. This process has the required low capacitance and low ON resistance for RF, and can interface with standard digital logic when combined with CMOS drivers in the same device. With best performance at frequencies below about 100 MHz, DMOS is useful up to about 1 GHz, with enhancements coming rapidly.

There are two DMOS processes, vertical and lateral. For RF applications, the lateral DMOS process (Figure 1, page 38) produces the best switches. Feedthrough capacitance is typically around 0.3 pF and channel ON resistance can be as low as 20 ohms, with 50 ohms typical for most configurations. With properly designed packaging, a DMOS switch can switch RF to 200 MHz with adequate isolation in the OFF mode. Incorporating compatible CMOS driving logic into the same device offers the convenience of control by conventional logic circuits. All of these factors make DMOS RF switches worth examination for new designs.

Applications Hints

There are a few aspects of DMOS switch characteristics that require special attention. These include harmonic distortion, power supply decoupling, and supply voltage levels.

Harmonic distortion is generated in an analog switch by the modulation of the channel ON resistance by the analog signal. Since the switch is controlled by the gate-source voltage, it is easy to see how the analog signal swing changes the

The Topaz CDG2214N

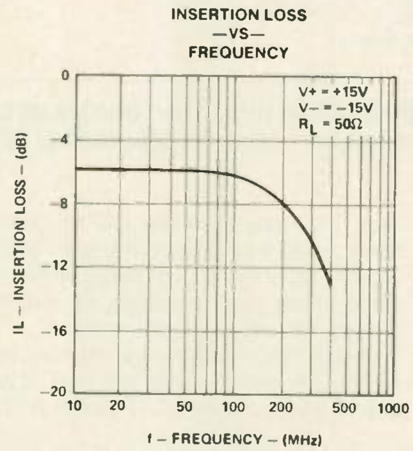

Figure 2. CDG2214N insertion loss vs. frequency.

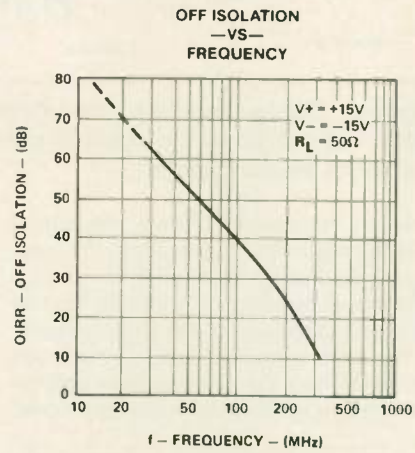

Figure 3. CDG2214N OFF isolation vs. frequency.

This is the first published description of a new analog RF switch from Topaz Semiconductor, the CDG2214N. A leader in lateral DMOS fabrication techniques, Topaz has introduced this device, a SPST switch with integral CMOS logic driver. The switch has been characterized for insertion loss and OFF isolation to beyond 200 MHz (Figures 2 and 3).

The CDG2214N is packaged in an 8-pin DIP package, with a single switch per package in a 50 ohm system, the switch has a typical insertion loss of 6 dB, increasing somewhat at frequencies above 100 MHz. OFF isolation falls off with frequency from 80 dB or so at low MHz ranges to 25 dB at 200 MHz. The device can handle analog signal swings of ±10 volts with ±15 volt supplies.

The integral CMOS driver operates from standard 5 volt, CMOS compatible logic input. Switching times are typically t_{on} = 50 nS and t_{off} = 20 nS, and are dependent mainly on the speed of the CMOS driver, not the DMOS switch element. Figure 4 shows the switching time test waveforms.

Finally, a significant attraction of solid state switching is price. The

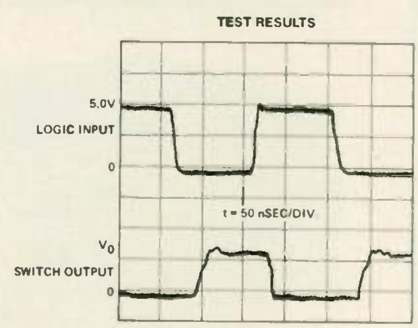

Figure 4. Switching time plot, showing analog switch output relative to the control logic input. The CMOS logic driver is the primary contributor to switching speed, as it is slower than the DMOS switch itself.

CDG2214N is offered at 99¢ each in 1,000 piece quantities. The CMOS/DMOS combination can be manufactured with low enough cost to make analog RF switching attractive for applications where it has never been used before. **Topaz Semiconductor, San Jose, Calif. For more information, circle INFO/CARD #214.**

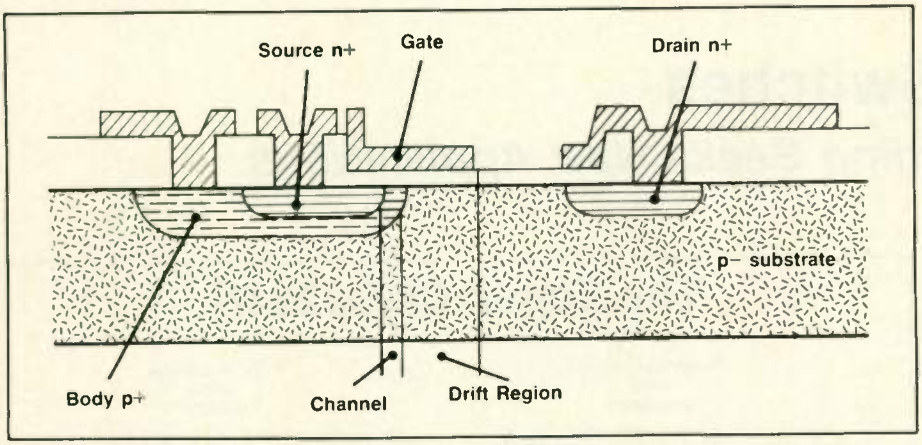

Figure 1. Cross-section of the lateral DMOS structure. The "double diffusions" of the body and source create the short channel length that is a key to its performance.

channel resistance. When the signal reaches the output termination, the varying resistance is converted into a changing voltage, resulting in (primarily) second harmonic distortion of the analog signal.

The main method of keeping distortion low is to keep the power supply voltage five volts higher than the maximum signal level. This will keep the ON resistance fairly constant by keeping the gate-source voltage out of the 2-5 volt range where the slope of the control voltage vs. channel resistance curve is quite steep.

Higher impedance can improve harmonic distortion, but at the cost of frequency response and OFF isolation. The OFF isolation can be improved with the TEE configuration, but frequency response will still suffer, due to increased capacitance and higher impedance. It should be noted that distortion is not as severe at higher frequencies, where the source/drain to gate capacitance also modulates the channel resistance, but counter to the purely voltage effects, reducing the overall distortion.

Power supply decoupling is essential for best performance. Logic signals in the control circuitry can introduce high frequency noise onto the power supply rails. Good decoupling methods at the power supply pins will keep the noise out of the device, where it could couple into the analog circuitry.

Although not perfect (is any device perfect?), lateral DMOS analog switches are a recent improvement in low power RF switching that should interest RF engineers. The technology is also producing amplifier transistors from small-signal to medium power levels. The low capacitance of the process makes sub-nanosecond switching possible, adding speed to the switching characteristics. It's a technology to keep an eye on. rf

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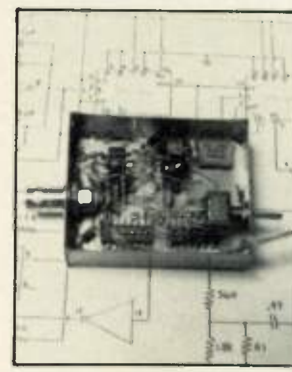

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RULES AND JUDGING CRITERIA

Judging Criteria

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1. Entries shall be RF circuits containing no more than 6 single active devices (tubes or transistors), or 4 integrated circuits, or be passive circuits of comparable complexity.

2. The circuit must have an obvious RF function and operate in the below-2 GHz frequency range.

3. Circuits must be the original work of the entrant.

4. If developed as part of the entrant's employment, entries must have the employer's approval for submission.

5. Components used must be generally available, not obsolete or proprietary.

6. Submission of an entry implies permission for *RF Design* to publish the material. If published, the entrant will receive the normal per-page author's honorarium.

7. Deadline for entries: **March 31, 1987.**

rf cover story

New Directions for Hewlett-Packard

HP Introduces Low-Priced Portable Spectrum Analyzer

In a significant departure from previous spectrum-analyzer offerings, Hewlett-Packard Company announces its first portable, low-priced RF spectrum analyzer, featured on this month's cover. The first public display of the unit will be at RF Expo East, Nov. 10-12, 1986, Boston, Mass.

The HP 8590A spectrum analyzer (10 kHz-1,500 MHz) weighs 29.8 pounds (13.5 kg), is sturdy and compact (21.3 cm × 36.6 cm×46.0 cm), and has a tilt-bail handle for easy carrying. At $9,500, the new fully-programmable instrument is priced substantially lower than previous HP spectrum analyzers. It has nearly all the data-processing and measurement functions of the HP 8568B high performance RF spectrum analyzer that sells for $34,600.

Intended for bench applications in R&D and manufacturing, and for on-site measurements at remote locations such as antenna sites, the HP 8590A has an optional cover that protects the front panel from dust, moisture and impact. This cover also provides storage for the operating guide or a handheld computer.

Simplified Manual Operation

The HP 8590A is the first HP spectrum analyzer to have both dedicated (single-function) pushbuttons and menu-labeled softkeys for manual operation. Frequently-used functions such as center frequency are activated via dedicated pushbuttons, while more than 80 additional functions are accessed by six softkeys labeled with CRT menus. The company believes this combination of softkeys and dedicated keys is the optimum design for easy manual operation. Most measurements are made using only three main controls on the front panel: FREQUENCY, SPAN and AMPLITUDE. The analyzer automatically adjusts internal parameters such as resolution bandwidths, sweep time, IF gain and input attenuation for an optimum CRT display.

Program with a Handheld or Desktop

Three different digital interface options are available for the HP 8590A: HP-IB (IEEE-488), HP-IL (Hewlett-Packard Interface Loop), and RS-232-C. This means the new analyzer can be programmed from a wide variety of computers, including HP technical workstations, personal computers such as HP Vectra PC and IBM PC, and handheld computers such as the HP-71B. The HP-71B computer can fit into the analyzer's optional front-panel cover for ease of carrying. With measurement programs written in BASIC and stored in the memory of the HP-71B, tests can be performed automatically at remote sites.

By bringing along a portable printer, the user can obtain report-quality test documentation at the test site.

Built-In Digital Functions

The HP 8590A has more than 100 digital functions to aid in measurement and data handling. The SIGNAL TRACK function holds a drifting signal at center screen and allows automatic "zooming-in" on a signal — that is, a fast reduction in frequency span and resolution bandwidth to permit close-in analysis.

Marker functions such as PEAK SEARCH and NEXT PEAK help find signals of interest, and MARKER DELTA measures amplitude and frequency differences between signals. It also is possible to stop the sweep at a marker position for up to 100 seconds, while monitoring the demodulated signal at the VIDEO OUTPUT.

Complex measurements such as percent AM, 99 percent power bandwidth and signal-to-noise are carried out automatically by other functions. A bandpass filter's bandwidth can be measured by 3 dB points or 6 dB points. The FFT function measures low-frequency AM sidebands such as 60 Hz hum, which are too low in frequency to be resolved by the analyzer's narrowest IF bandwidth.

Digital-trace functions include two active traces, trace storage, MAX HOLD (retains maximum trace values) and VIDEO AVERAGE (smooths displayed-noise peaks to help in locating low-level signals). Trace-math functions allow frequency response variations caused by an external source to be cancelled out (normalized) in stimulus-response testing.

Other functions allow trace data to be output directly (via the appropriate digital interface) to a plotter or printer, permit up to nine different sets of analyzer-control settings to be stored for later recall, and automatically calibrate the analyzer for accurate amplitude and frequency measurements.

RF Performance

The HP 8590A has Gaussian-shaped resolution bandwidths from 1 kHz to 3 MHz in a 1, 3 sequence and video bandwidths from 30 Hz to 3 MHz. Frequency-readout accuracy is ±5 MHz (+1 percent of span) and frequency response is ±1.0 dB. Maximum sensitivity is −115 dB in a 1 kHz IF bandwidth from 100 kHz to 1.2 GHz.

The HP 8590A portable RF spectrum analyzer is $9,500. Each digital-interface option is $450. The optional front-panel cover is $200. For further information circle INFO/CARD #129.

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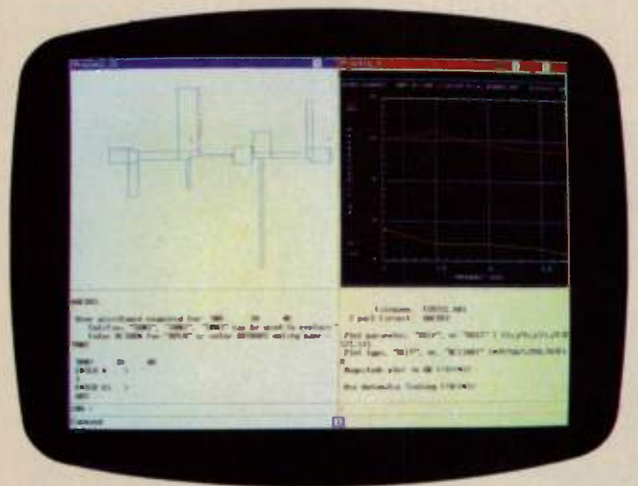

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Antennas for EMI Measurements

Part I: Measurement Requirements and Antenna Characteristics

By Edwin L. Bronaugh
Electro-Metrics
Division of Penril Corporation

Measurement of radiated electromagnetic interference (EMI) seems simple at first glance, and exceptionally complex upon further examination. The concept of measuring field intensity in volts/meter is straightforward, but the practice of achieving accurate measurements requires both the art and science aspects of engineering. The author has prepared a comprehensive examination of radiated EMI measurement, presented in two parts. This month, the basic theory of EMI measurement and the related characteristics of antennas in general are presented. In December, the second part will describe the performance characteristics of specific antenna types used for radiated EMI measurements.

This article is about radiated EMI measurements, but the term *radiated* if often a misnomer since the measurements are usually made under conditions which do not permit actual measurement of the power being radiated from the EMI source. Radiated power (true radiation or Poynting vector) measurements are properly made in free space at great distances from the source and any reflectors or other field perturbing objects. Radiated EMI measurements are often made in the near field zone of the source, usually near the ground, and in the near field zone of the measurement antenna in many parts of the frequency spectrum. A large body of measurements, those made in accordance with military standards, are almost always made in shielded enclosures where multiple reflections exist. The term *radiated* is used in EMI measurements to signify field strength measurements; electric-field (E-field) strength measurements and magnetic field (H-field and B-field) strength measurements. Under some conditions plane-wave field strength measurements are said to be made, but these are often made using E-field responding antennas such as dipole, dipole-array, or monopole-fed-horn antennas so it is usually better to think of all radiated EMI measurements as being either E-field or H-field (B-field) strength measurements.

The fields being measured are seldom plane, homogeneous, or uniform. The field lines are often curved, the vectors of the electric and magnetic components are often out of phase, and the waves that exist are often ellipsoidally polarized (10). Antennas for measuring these fields, then, must have characteristics which will provide the truest feasible response to the strengths of the fields, and the rules for using the antennas must assure that they are used as intended.

In specifying the characteristics of an antenna for EMI measurements, the wavelength of the EMI to be measured, relative to the dimensions of the test setup and of the antenna, is one of the major factors. Radiated EMI tests may require a distance of separation ranging from one meter for military standards to as much as 300 meters for some civil standards, with 3 to 30 meters being typical. These measurement distances correspond to one wavelength at frequencies ranging from 300 MHz down to 1 MHz. Practical dimensions for antennas to be used in shielded enclosures are about 2.5 meters, maximum, and for antennas to be used on open area test sites may reach six meters; but such large antennas are cumbersome to use and will likely give poor results.

What does this mean? It means that at frequencies below about 30 MHz, the measurements are being made in the near field regions of both the source of EMI and the antenna. The selection of 30 MHz is somewhat arbitrary, but coincides rather well with the characteristics of many sources of EMI. Between 30 and 100 MHz, the source is usually in a transition zone between the near field and far field regions of the antenna, while the antenna may be in the near field region, the transitional zone, or the far field region of the source, depending upon the source dimensions and radiation characteristics. Above 100 MHz, the source is usually in the far field region of the antenna, if the antenna gain is not too high, but the antenna may still not be in the far field region of the source.

In deciding what measurement to make and thus what the antenna characteristics should be, the tacit assumption is that above 20 to 30 MHz the E-Field strength behaves much as would be expected if the measurement point were in the far field region of the source, and thus the propagation of EMI from the source to a more distant receptor (susceptor) can be reasonably predicted on the basis of the E-Field strength alone. Below about 20 to 30 MHz, both E-Field and H-Field strengths are often measured at the lower frequencies, the H-Field is usually chosen because it is better behaved, easier to measure, and for many sources can be used for practical prediction of the EMI field strength at distant receptors.

Required Antenna Characteristics

The H-Field below 30 MHz is easiest to measure. An unambiguous measurement can be made with an electrically small loop antenna which does not respond to the electric field. An electrically small loop antenna must have a diameter much less than λ/π (λ is the wavelength). The smaller the loop is physically, the less it tends to average the spatial field strength variations, and thus the more accurate is the measure of the field strength at the measurement point. Another way to put it is that the smaller the loop, the more nearly planar is the field across the antenna, and the more accurate the measurement.

The E-Field below 30 MHz is more difficult to measure because the fields are perturbed by dielectric materials, thus the antenna itself, tripods, people and other objects tend to disturb the measurements. The E-Field is measured by an electric responding transducer such as an electrically short monopole or dipole antenna. To be considered electrically short, a monopole should be much shorter than $\lambda/4$ and a dipole should be much shorter than $\lambda/2$. The shorter the antenna, the more nearly planar the wave across it and thus the more accurate the measurement.

In the frequency range above 30 MHz, the antenna should either be electrically short, or the source of EMI should be in the far field region of the antenna.

For simple dipole antennas, the middle of the transition zone is marked by the distance $\lambda/2\pi$, and the transition zone occupies a shorter range of distances for electrically short antennas than for electrically large, e.g., half-wavelength dipole antennas. Half-wavelength dipole antennas should be used at a distance much more than $\lambda/2\pi$ from the source of EMI with a good "rule of thumb" being five times this distance. Electrically short dipole antennas should be more than their half-lengths from the source of EMI. Again, the smaller the antenna and/or the farther it is away from the source (or from the field-perturbing objects), the more accurate the measurements.

For complex arrays or aperture antennas (horns, etc.) having gains higher than simple dipole antennas, the far field region starts at a distance inversely related to the gain of the antenna. For aperture antennas, this distance is commonly accepted to be $2D^2/\lambda$, where D is the largest linear dimension of the aperture.

Wide-bandwidth antennas provide for much more convenience and efficiency in making EMI measurements. Antennas which must be readjusted at each frequency make measurements much more time consuming and tend to prevent their automation. For that reason, electrically small antennas, biconical dipole antennas, log periodic dipole array antennas, conical log spiral antennas, and horn antennas tend to be more popular for EMI measurements than half-wavelength tuned dipole antennas.

EMI Antenna Design

The study and understanding of the design and use of EMI antennas could start many places, but since the antenna factor relates the design of the antenna to its use through its performance we will start with the Antenna Factor: What it is, what factors it contains, how it is measured and how it is used.

The antenna factor for an EMI antenna is defined by equation (1) which shows the relationship between the loaded two-terminal output voltage V_0 of the antenna and the strength E of the electric field in which it is immersed. The units are volts and volts per meter, respectively.

$$AF = E/V_0 \qquad (1)$$

This deceptively simple factor may be analyzed as the product of several coefficient factors which describe mathematically how the antenna works. Each of the coefficients is the result of the functioning of some part of the antenna, so that the number of them included in the antenna factor and their values depend upon the design and construction of the antenna. Two of the coefficients which are always included are a *conversion factor* which converts from values of field strength to values of open-circuit antenna terminal voltage and a *load correction factor* which relates the voltage across the load circuit connected to the antenna to the open-circuit antenna terminal voltage. In addition to these, there are usually a number of other coefficients to correct for impedance mismatches, transformations, and losses in the several parts of the antenna.

The conversion factor is the kernel of the antenna factor, having within it the effective height or length of the antenna including gain and pattern effects. The conversion factor is usually given for the antenna pattern maximum since that is the way in which the antenna is typically used for EMI measurements. For dipole and monopole antennas, the conversion factor is the reciprocal of the effective length or height, for loop antennas it is proportional to the reciprocal of the square-root of the effective area. The gain and effective area of a complex antenna are alway related to power transfer. Thus, the square-root of these quantities is used when they are related to the electric field strength or the terminal voltage, as is done in the antenna factor.

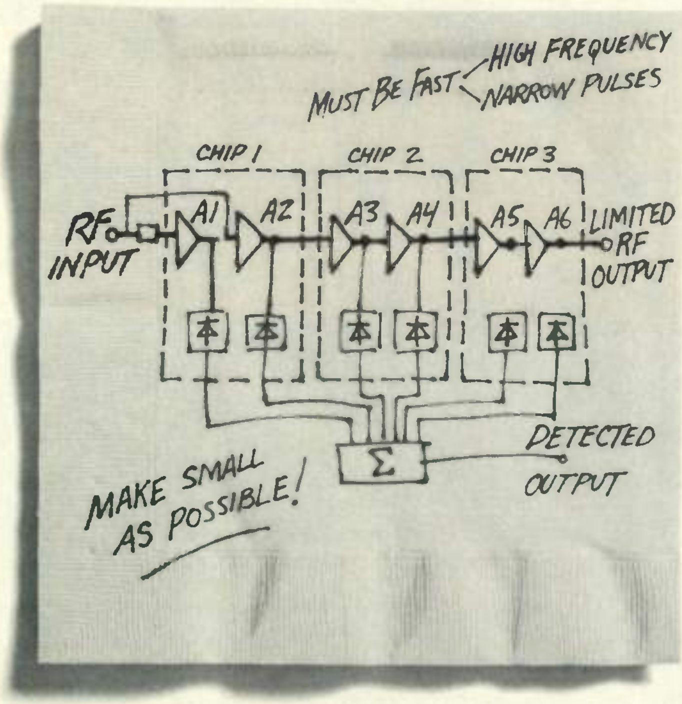
The load correction factor accounts for the ratio between the open-circuit voltage at the antenna terminals and the closed circuit voltage across the load connected to the antenna. For a relatively simple antenna such as resonant dipole, this is often taken to be a factor of two (6 dB). This is based upon the idea that the load impedance matches the antenna feed-point impedance, e.g., approximately 73 ohms for an ideal half-wavelength resonant dipole antenna. Any mismatch between the load and the antenna impedance is accounted for by a separate "mismatch" factor.

If no impedance matching is provided between the antenna and load, then a coefficient to account for the "mismatch loss" must be included in the antenna factor. This mismatch loss is not energy dissipated as in a resistance; rather it is energy reflected from a point of mismatch, the impedance discontinuity, back into the antenna to be reradiated.

Use of a BALUN

A balanced antenna such as a dipole must be kept balanced if it is to perform correctly. If it is connected to the EMI meter by a coaxial transmission line (as is commonly done), then it must contain a balanced-to-unbalanced transformer (BALUN). In some antennas this transformer may also perform an impedance transformation along with its balanced-to-unbalanced transformation. This BALUN may be realized as a lumped-element circuit such as a set of tightly coupled windings of wire on a ferrite toroid or as a distributed element circuit using pieces of transmission line. In any case, its transforming action and its losses must appear among the coefficients in the antenna factor, either implicitly or explicitly.

In some antennas, a carefully designed BALUN provides balancing for the antenna, but impedance matching is done by a resistive attenuator proportioned so that the antenna is matched, the transmission line is matched, and the entire loss is 10 dB. This makes the antenna factor larger, but both the antenna and its transmission line are properly terminated over a very wide band of frequencies. This



The best hybrids for radar aren't always hybrids.

They're SL2521 Monolithic Log Amps from Plessey. You heard right. Monolithic.

Our advanced 3 micron oxide isolated bipolar process allows us to put Phase, Video and IF limiting functions on a single chip. About one-third the size of bulky hybrids.

So now you can achieve a higher level of performance and reliability with less power than most hybrids. Not to mention a lower price.

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And, like Plessey's complete family of monolithic ICs, the SL2521 meets MIL-STD-883C requirements.

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So if you still think a radar hybrid has to be a hybrid, we'd like to change your mind. Call Plessey at 800-321-0871 or 714-951-5212. For free on-line information, dial 1-800-345-7335.* Or write us at Plessey Semiconductors, 3 Whatney, Irvine, CA 92718.

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*Use any 80-column ASCII terminal or PC and a 300 or 1200 baud modem (EVEN or IGNORE parity, 7 data bits, 1 stop bit.) At "Enter Response Code:" Type LOGAMP. In Conn. Dial (203) 852-9201.

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allows for a much more accurate calculation of the antenna factor of the antenna from its physical properties, but such an antenna will still perform inaccurately if it is used improperly.

The preceding discussion has given a description of the antenna factor, telling what it is and indicating how it might be found. However, except for certain antennas finding the antenna factors by calculation from design parameters is fraught with many potential errors. (The performance of antennas such as standard gain horns, and some electrically small electric and magnetic field sensors can be determined quite accurately and reliably as a function of their geometry.) For most antennas, it is far better to measure the antenna factors by the careful use of a technically sound procedure (1). There are four general approaches to making antenna factor measurements. In one, the gain of the antenna is measured under conditions that may simulate the actual use of the antenna. In the second, a standard field is produced and the antenna factor is determined directly by placing the antenna in the field and calculating the ratio of the field strength to the measured output voltage. In the third method, the antenna to be calibrated is substituted in a constant field for an antenna of known performance and the antenna factor calculated by comparison. In the fourth method, the attenuation between three pairings of three antennas is measured on a standard test site and the antenna factors for each antenna calculated from the proven propagation theory for the site. These methods are discussed in more detail in references (1, 2, 3 and 4).

The use of the antenna factor is readily apparent, but not so easily seen are the implications of its use. Referring to equation (1), note that the units of the antenna factor are reciprocal meter. When the antenna factor is multiplied by the two-terminal voltage measured at the output of the antenna, this voltage is converted from volts to volts (or amperes) per meter, in terms of field strength. One can see that the strength of an electromagnetic field may be accurately measured by any antenna for which the antenna factor is accurately known. Usually the antenna factor is given in dB (m^{-1}) or dB(S/m) and it is added to the two-terminal antenna output voltage in dB(uV) measured by an EMI analyzer giving the field strength in dB(uV/m) or dB(uA/m).

Since factors are usually given and used in decibels, it will help to first restate

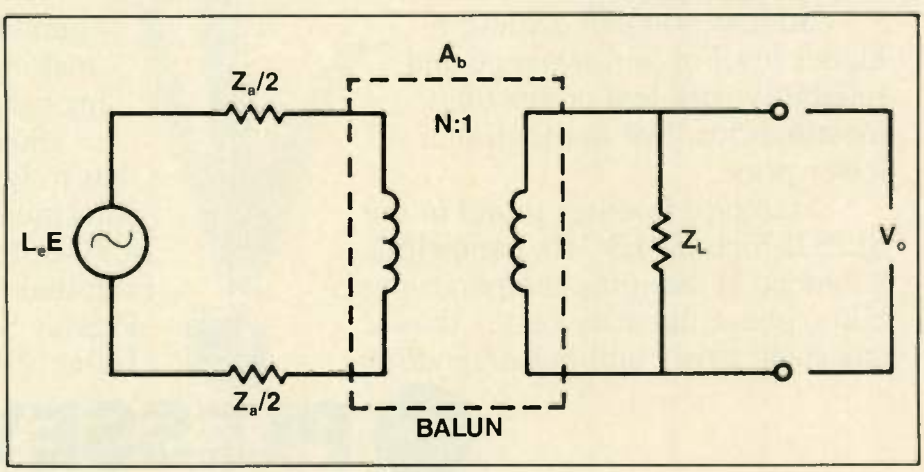

Figure 1. Simplified circuit model for a dipole antenna.

equation (1) in dB as shown in equation (2). In the remainder of this discussion, we will be considering the antenna factor for electric field or planewave antennas, but the ideas also apply to magnetic field antennas.

$$AF = 20 \log E - 20 \log V_o \quad (2)$$

The antenna factor is in dB(m^{-1}). The values of 20 log E and 20 log V_o are usually in dB(uV/m) and dB(uV). Common logarithms are used. A more specific relationship for dipole antenna factors is shown in equation (3).

$$AF = 6 - 20 \log L_{ant} + A_o + 20 \log [(Z_{ant} + N^2 Z_L)/2NZ_L] \quad (3)$$

Figure 1 shows the circuit model of the dipole antenna used in this equation. The symbols used in equation (3) and Figure 1 have the following meanings:

The constant 6 is the 6 dB conversion from open circuit voltage to the voltage across a matched load.

L_{ant} is the effective length of the antenna in meters.

Z_{ant} is the impedance of the antenna as a source (feed-point impedance) in ohms (Z_{ant} may be resistive, or complex).

Z_L is the impedance of the load connected to the antenna in ohms.

A_b is the loss (not including the effects of the transformation ratio, if any) in the BALUN in dB.

N is the antenna-to-load voltage transformation ratio of the BALUN (usually either unity, N = 1, or scaled for impedance matching, $N = \sqrt{Z_{ant}/Z_L}$).

In the general case, the impedances are complex numbers, i.e., $Z_{ant} = R_{ant} + jX_{ant}$ and $Z_L = R_L + jX_L$, and the BALUN does not provide a perfectly resistive transformation. These effects are usually ignored when the antenna factor is calculated on the basis of "theoretical" values, because they make the calculation much more cumbersome if included. However, ignoring them causes the calculated antenna factor to be incorrect. These concepts will be used or referred to next month in Part II: Characteristics of Specific EMI Antennas. rf

References

1. Bronaugh, "Open-Site Verification of EMC Antenna Factors," *ELECTRO/83 Record,* Session B, New York, New York, April 19-21.

2. SAE ARP 958-1968, "Broadband Electromagnetic Interference Measurement Antennas; Standard Calibration Requirements and Methods," SAE, 400 Commonwealth Drive, Warrendale, Pennsylvania 15096.

3. F.M. Greene, "NBS Field-Strength Standards and Measurements (30 Hz to 1000 MHz)," *Proc. IEEE,* Vol. 55, No. 6, pp.970-981, June, 1967.

4. A.A. Smith, Jr., "Standard-Site Method for Determining Antenna Factors," *IEEE Transactions on Electromagnetic Compatibility,* Vol., EMC-25, No. 3, pp.316-322, August 1982.

10. T. Dvorak and G. Meyer, "Influence of Ground Reflection in Open Site Measurements of Broadband Interference," *1979 IEEE International Symposium on Electromagnetic Compatibility Record,* 79CH1383-9EMC, October 9-11, 1979, San Diego, California, pp.369-374.

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Edwin L. Bronaugh is Director of Research and Development at Electro-Metrics, a division of Penril Corporation, 100 Church Street, Amsterdam, N.Y. 12010. Mr. Bronaugh is recognized as a leading authority on electromagnetic compatibility (EMC), and is a member of several EMC standards committees.

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Low Cost Monolithic

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These SPST, SPDT and SPMT switches combine broad bandwidth performance with high isolation and low insertion loss. The inherent isolation in the dc bias lines eliminate the need for dc blocks and rf chokes, thus extending the operational frequency range.

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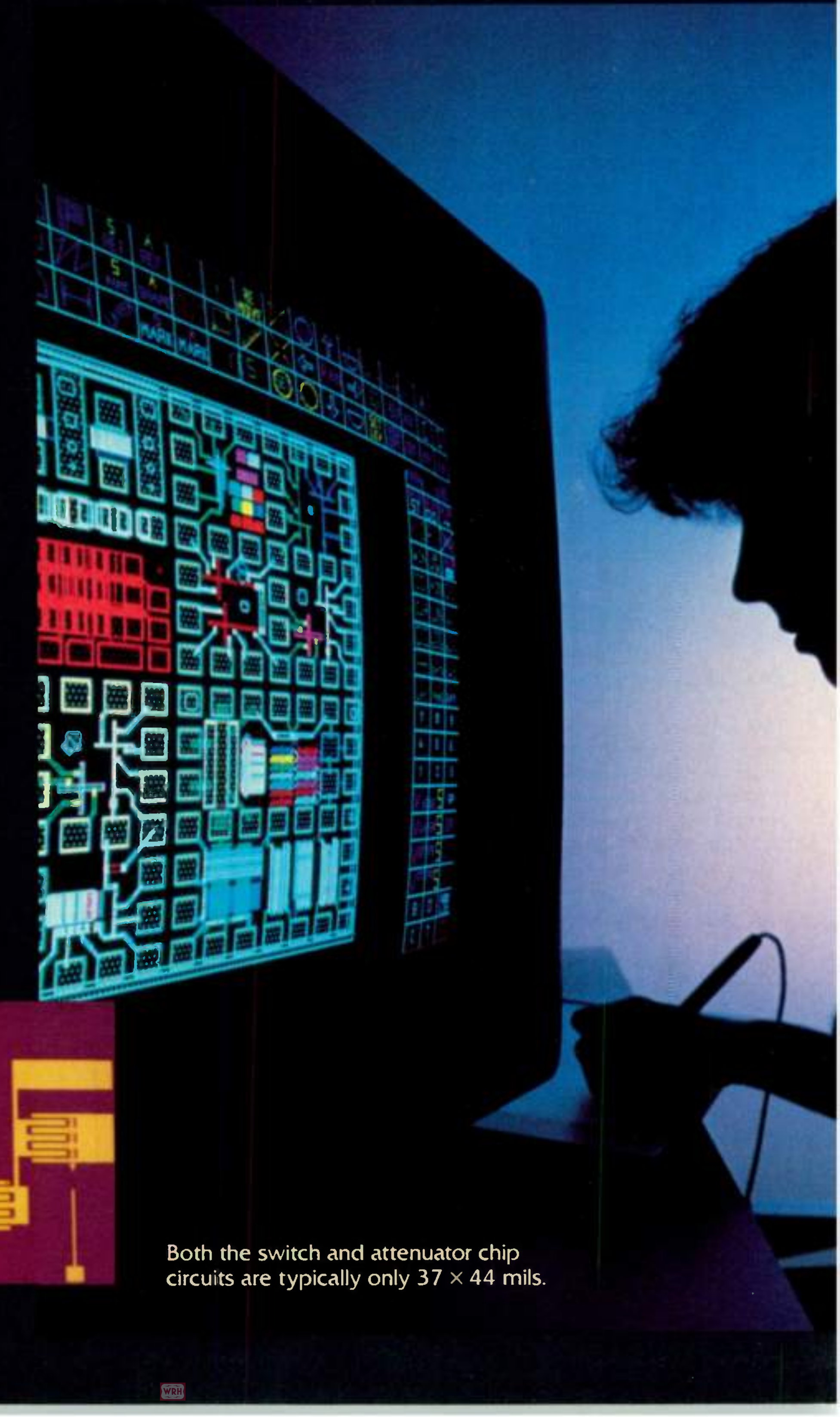

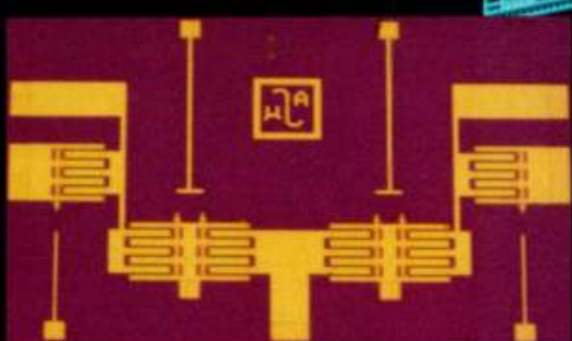

Both the switch and attenuator chip circuits are typically only 37 × 44 mils.

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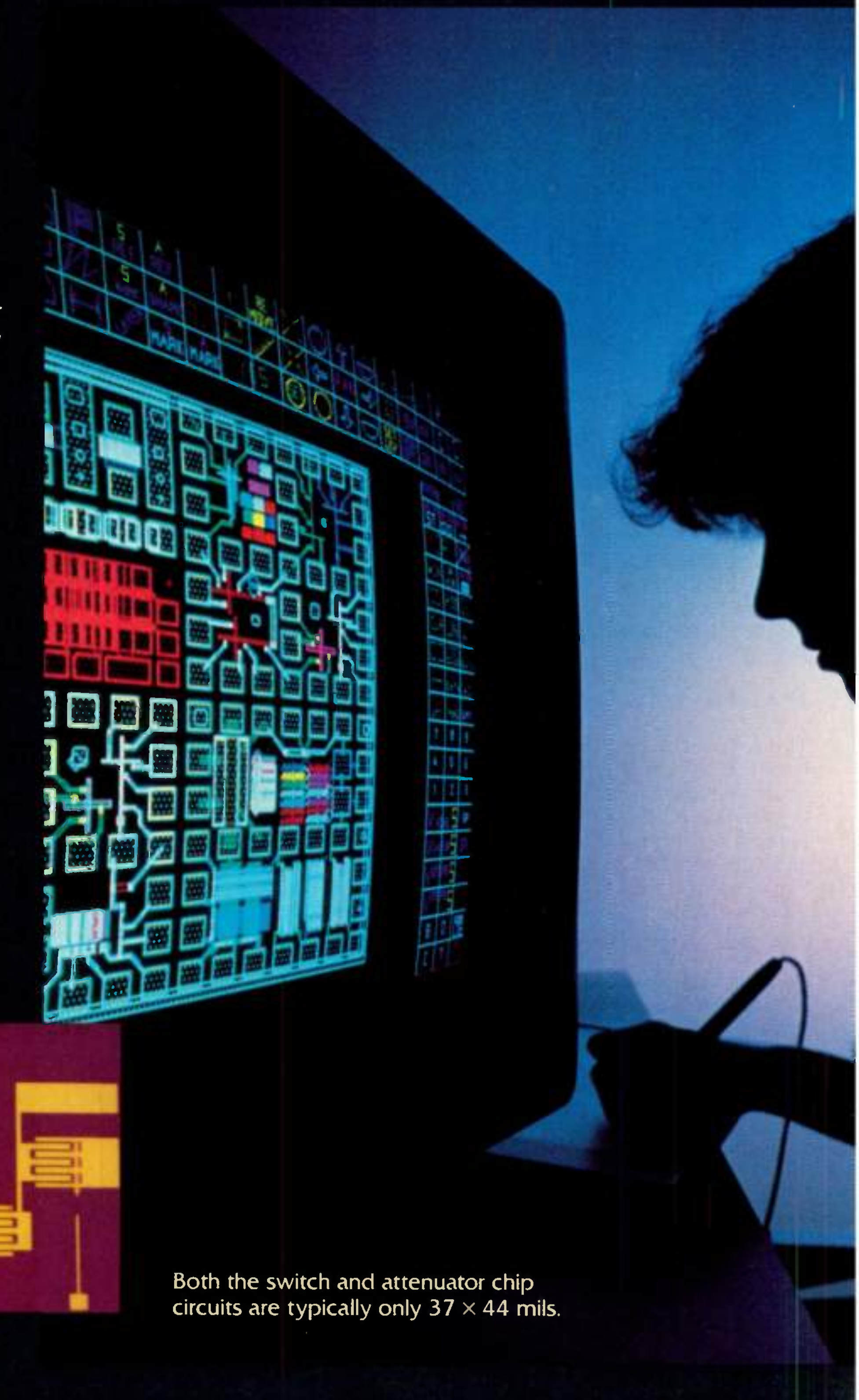

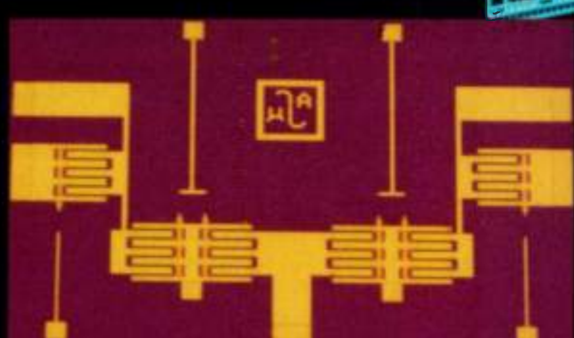

Both the switch and attenuator chip circuits are typically only 37 × 44 mils.

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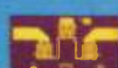
GaAs Control Circuits

ASCADABLE CONTROL CIRCUITS dc TO 20.0 GHz

SWITCH MODELS	TYPICAL PERFORMANCE (T_a = 25°C)				CIRCUIT CHIP
	FREQ	INSERTION LOSS (dB)	ISOLATION (dB)	VSWR	
MA4GM 201 (SPST)	f_0 = 500MHz	0.28	52	1.1	
	f_0 = 2GHz	0.30	40	1.2	
MA4GM 202 (SPDT)	f_0 = 500MHz	0.60	62	1.3	
	f_0 = 2GHz	0.85	45	1.6	
MA4GM 211 (SPST)	f_0 = 5GHz	1.7	28	1.8	
	f_0 = 12GHz	2.5	17	2.0	
MA4GM 212 (SPDT)	f_0 = 5GHz	1.6	20	1.8	
	f_0 = 12GHz	2.7	16	2.0	
MA4GM 221 (SPST)	f_0 = 10GHz	1.3	15	1.7	
	f_0 = 18GHz	1.7	10	1.5	
MA4GM 222 (SPDT)	f_0 = 10GHz	1.5	23	1.4	
	f_0 = 18GHz	2.6	14	1.6	

ATTENUATOR MODELS	TYPICAL PERFORMANCE (T_a = 25°C)				CIRCUIT CHIP
	FREQ	ATTENUATION RANGE (dB)	INSERTION LOSS (dB)	VSWR	
MA4GM 301	f_0 = 500MHz	25	1.2	1.4	
	f_0 = 2GHz	20	1.3	1.5	
MA4GM 311	f_0 = 5GHz	14	1.2	1.1	
	f_0 = 12GHz	12	1.3	1.4	
MA4GM 321	f_0 = 10GHz	11	1.3	1.4	
	f_0 = 20GHz	8	1.8	1.7	

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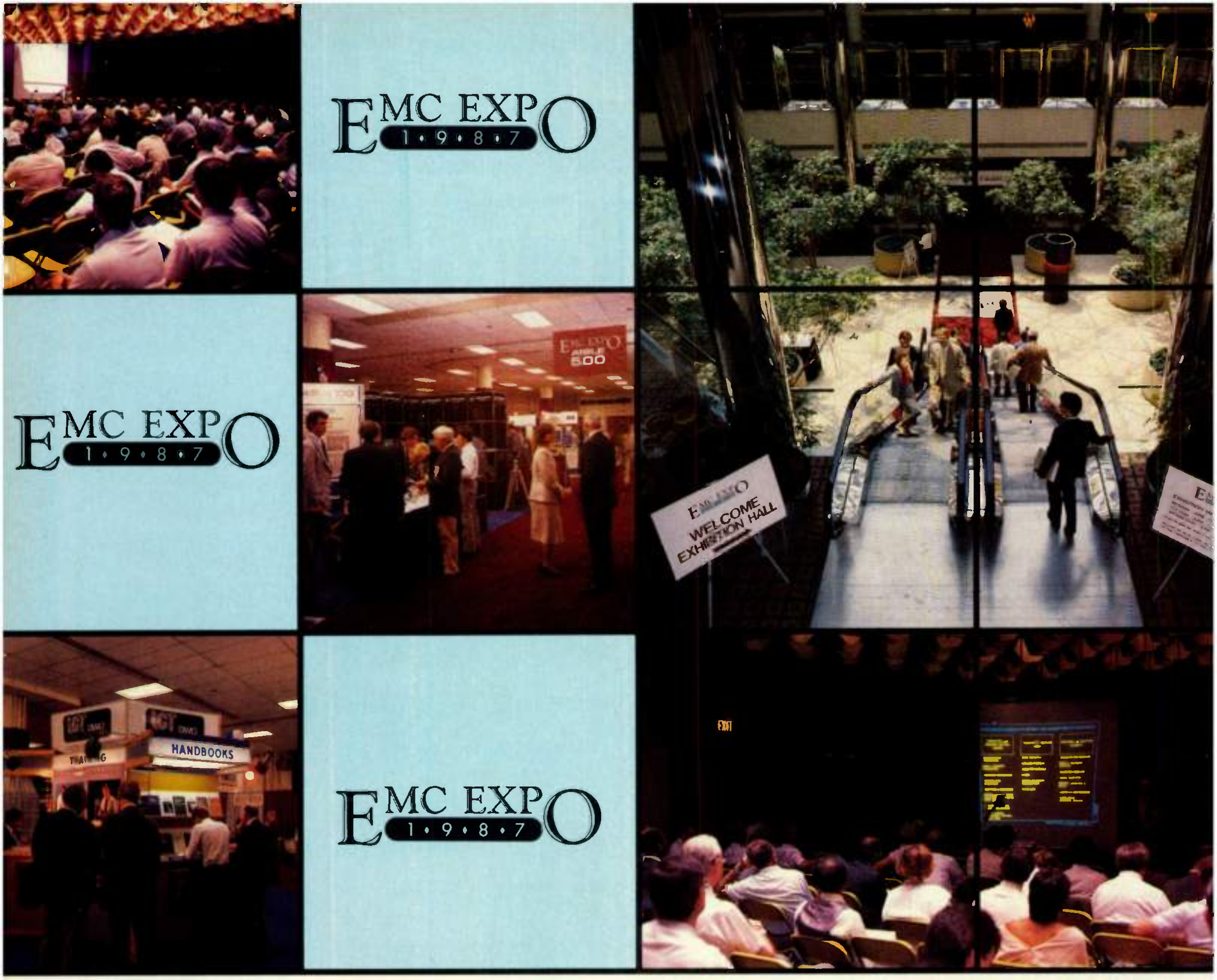
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Hewlett-Packard's Spokane Division Presents High Performance Phase Noise Measurement System.

The HP 3048A is a new, high performance phase noise measurement system with a noise floor low enough to measure even very low noise reference sources. This means that the actual noise sensitivity is limited only by the reference source used to demodulate the phase noise. Measurement by the HP 3048 over the 0.01 Hz to 1 MHz offset range is specified to an accuracy of ±2 dB. At offsets greater than 1 MHz, accuracy is specified at ±4 dB, to accommodate the variety of spectrum analyzers that the system can control to measure this range.

The HP 3048A system includes the HP 11848A Phase Noise Interface, including the necessary phase detectors, amplifiers, filters and switches for phase noise measurement over 5 MHz to 18 GHz. An input for an external phase detector for other frequencies is provided. The HP 3561 Dynamic Signal Analyzer, another part of the system, makes Fourier-type measurements on signals within a μHz to 100 kHz. Built-in averaging and large dynamic range make it capable of quantifying demodulated phase noise.

The final part of the basic system is the System Software, providing the measurement program and the BASIC operating system for the controller.

Options include adding the HP 8662A, HP 8663A, or the HP 8642A or B model signal generators, the HP 3585A spectrum analyzer, or the HP 11729C carrier noise test set. AM noise detector, high frequency phase detector, and a system rack are other possible options. Arrangements can be made to include compatible equipment that may already be available to the user.

The basic system with HP 11848A, HP 3561A and operator's training course (on-site training) has a price of $31,500. **Hewlett-Packard, Spokane Division, Spokane, Wash. INFO/CARD #177.**

Synthesized Signal Generator: John Fluke Mfg. Co.

The 6061A programmable synthesized signal generator is the latest addition to the 6060 family, with increased spectral purity and performance features. The 6061A has a frequency range of .01 to 1050 MHz with 10 Hz resolution. Amplitude range is from −127 to +13 dBm with 0.1 dB resolution and an absolute accuracy of ±1 dB. Internal and external AM and FM can be used in combination or separately. U.S. list price is $5695. A high stability reference oscillator option (−130) is $850, medium stability reference option (−132) is $300, and a rear panel RF output option (−830) is $105. **John Fluke Mfg. Co., Inc., Everett, Wash. Please circle INFO/CARD #176.**

2-8 GHz Drop-In Amps: Microwave Research & Mfg., Inc.

MRA-8 drop-in amplifiers will be introduced by MRM™. With 15 dB gain, these devices provide broadband amplification over the 2-8 GHz range in a microstrip-compatible package. **Microwave Research & Mfg., Inc., Hopedale, Mass. INFO/CARD #175.**

Low Noise, Low Cost GaAs FET: Motorola Semiconductor Prod. Inc.

Motorola has re-entered the GaAs low noise marketplace with the introduction of a low cost field effect transistor. The new MRF966 dual-gate FET provides lower gate leakage (1 μA typical), higher gm (18 ms minimum), higher gain than bipolar (17 dB typical at 1 GHz), and low noise (1.2 dB typical at 1 GHz). The MRF966 is a mass-producible, ultra reliable FET available in a plastic macro-X package. The chip (MRFC966) is also available. Pricing for the MRF966 is $1.95 (100s) and below $1.00 in volume quantities. **Motorola Semiconductor Products Inc., Phoenix, Ariz. INFO/CARD #174.**

Controllable Power Amplifier: ENI, Inc.

The ENI solid state amplifier Model AP-400 is the company's first broadband amplifier offering a controllable output power. The amplifier covers the frequency range of 80 kHz to 2.7 MHz with a maximum power of 400 watts. The unit may be operated either as a conventional amplifier in the unleveled mode, providing 55 dB of gain, or as a controllable power source in the leveled mode. The power level adjustment is either by front panel control or a 0 to 10 volt DC signal. The built-in power meter provides an accurate indication of both the forward and load power. A rear panel connector allows the amplifier to be externally controlled and monitored. Protected against excessive mismatch by output power limiter. The AP-400 is priced at $3,250. **ENI, Inc., Rochester, N.Y. INFO/CARD #173.**

TFE Capacitors and Substrates: Polyflon Company

Cuflon® is an ultra low loss, high-Q microwave substrate, featuring a pure TFE dielectric with a dielectric constant of 2.1. Cuflon has been successfully used in applications up to 100 GHz. Copper cladding is electroplated directly to the TFE, eliminating added loss and instability of a secondary dielectric adhesive. Polyflon will also have on display its new models of TFE-insulated non-magnetic fixed and variable RF capacitors for applications requiring high Q, 0.1-1000 pF, and voltages up to 50 kV. **Polyflon Company, New Rochelle, N.Y. Please circle INFO/CARD #172.**

Wideband Image-Reject Mixers: Merrimac Industries, Inc.

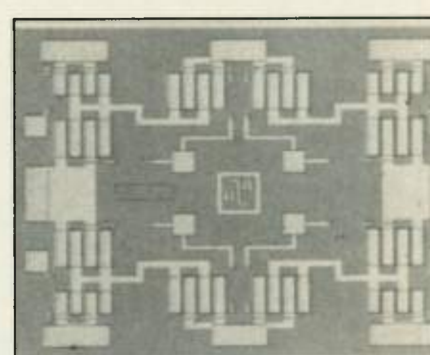
Merrimac introduces a new line of image-reject mixers featuring multi-octave bandwidths. The IRMS series covers RF/LO frequencies from as low as 1 MHz to over 250 MHz for an IF output in the range of 10 MHz to 150 MHz. Typical image rejection is in excess of 25 dB, with a carrier suppression of 30 dB and a conversion loss of less than 10 dB. Tighter specifications can be met for specific frequencies and bandwidths. The IRMS series is priced from $500.00. **Merrimac Industries, Inc., West Caldwell, N.J. INFO/CARD #165.**

400 MHz Silicon Op Amp: Plessey Semiconductors

Plessey Semiconductors introduces a monolithic bipolar high speed, high performance operational amplifier with a unity gain bandwidth of 400 MHz. Designated the SL9999, the new part is designed as an ideal companion for high speed flash analog-to-digital converters, but is appropriate for other circuits including those with low impedance and high capacitive load. The integral DC buffer and output DC offset circuitry provide flexibility for use as wideband buffer/level shifters, wideband IF amplification, fast settling pulse amplifiers, video amplifier/line driver, and other high speed op-amp applications. The SL9999 is priced at $13.20 (1000s) in a 16-pin dual in-line ceramic package. A leadless chip carrier version will be available soon. **Plessey Semiconductors, Irvine, Calif. INFO/CARD #171.**

MMIC Multi-Function Controls: M/A-Com Advanced Semiconductor

New GaAs FET MMIC dual function switches have been developed, combining a transfer switch function with a variable absorptive attenuator. These multi-function operations can be achieved with this single MMIC chip. The circuit is



offered in two configurations: a 0.5 watt unit, P/N MA4GM316 and a 0.25 watt unit, P/N MA4GM315. These circuits have an operating band of DC to 12 GHz, and are designed to complement M/A-Com's existing MMIC control circuit product line. **M/A-Com Advanced Semiconductor Operations, Lowell, Mass. Please circle INFO/CARD #170.**

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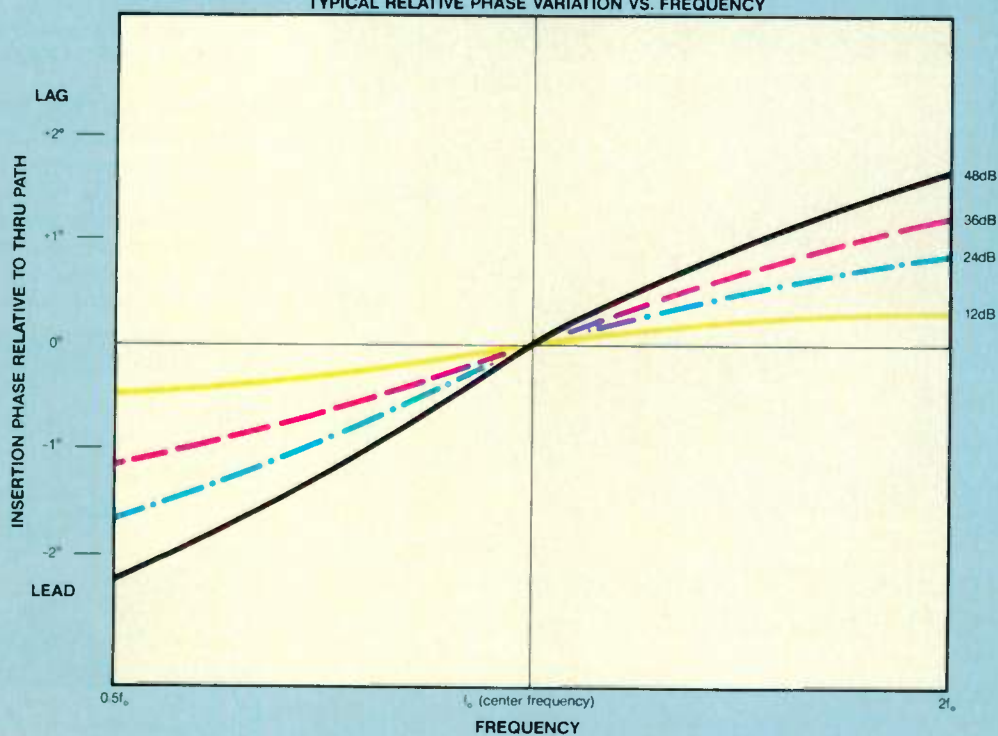

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Low Cost Voltmeter: Boonton Electronics Corp.

The low cost Model 92EA RMS Voltmeter, option −16, features true rms voltage measurements from 200 μV to 3 V over a frequency range from 10 Hz to 20 MHz. High input impedance, excellent stability, and low noise are provided by a passive (non-sampling) probe. Adapters are available for through-line or terminated measurements in both 50Ω and 75Ω systems.

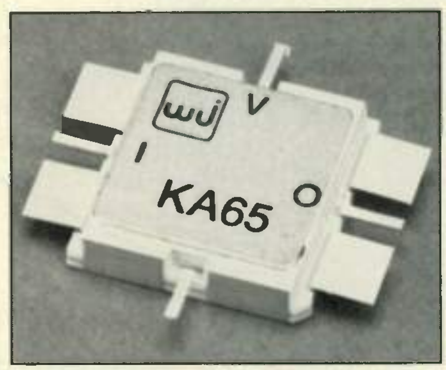
The 93EA has two linear voltage scales and a dBm scale referred to 1 mW in 50Ω. A dBm scale referred to 75Ω is an option, as are dBV and dBmV scales. Price of the Model 92EA, option −16, is $1,845. **Boonton Electronics Corporation, Randolph, N.J. INFO/CARD #169.**

2.6 GHz Frequency Counter: Racal Dana Instruments Inc.

The new model 1999 frequency counter is introduced by Racal-Dana, offering nine-digit resolution in one second from 10 Hz to 2.6 GHz. An external arming input allows measurements on signal bursts, or for segments of a waveform. Also included is a single-shot and hold capability. Single-key nulling allows measurements to be made relative to a previous measurement, for offset or drift measurements. Options available include GP-IB and high stability reference oscillators. **Racal-Dana Instruments Inc., Irvine, Calif. INFO/CARD #168.**

Two New 2-6 GHz Amplifiers: Watkins-Johnson Company

New introductions from Watkins-Johnson include the KA62 and KA65 GaAs FET amplifier modules in W-J's type "KA" microstrip-compatible package. The KA62 has specifications of: 15 dB gain, +12.5 dBm output power, 4.0 dB typical noise figure and 2.1:1 VSWR. The KA62 operates from a +5 volt supply. The KA65 is a +12 volt device, with 14.5 dB gain, +19.0 dBm output power, 4.5 dB typical noise figure and 2.1:1 VSWR. Both devices are specified over the temperature range from −54 to +85 degrees Celsius. **Watkins-Johnson Company, Palo Alto, Calif. INFO/CARD #167.**



New Release of CIAO Software: SPEFCO Software

CIAO (Circuit Analysis & Optimization) is a program operating in the frequency domain for one-port or two-port networks containing resistors, capacitors, inductors (coupled), transmission lines and stubs, controlled sources, and one-, two- or three-ports described by tables of S, Y, or Z parameters. New features in the latest release include parallel-coupled microstrip lines and directional couplers, plus user-definable models. Complete instructions and examples are given for the creation of models by the user. SPEFCO also announces international representation, with agencies in England and France. **SPEFCO Software, Stony Brook, N.Y. INFO/CARD #166.**

1 GHz Custom Signal Processor: KDI Electronics, Inc.

This custom microwave device is illustrative of the advances that have been made by KDI Electronics in the design and production of thick film microwave integrated circuits and sub-assemblies. The device contains PIN diode switches, attenuators, resistive power dividers and internal driver circuits for TTL compatibility. The entire circuit is fabricated using thick film technology. **KDI Electronics, Inc., Whippany, N.J. INFO/CARD #164.**

Low Noise, High Gain MMICs: Avantek, Inc.

Avantek has introduced three new DC to 3 GHz MODAMP™ cascadable silicon bipolar MMIC amplifiers which can operate with power supply voltages as low as +5 VDC. These MMICs are general purpose, cascadable gain blocks intended for use in narrow or broad bandwidth IF and RF amplifier design. At 1 GHz, models MSA-0735/-0770/-0885 typically feature 12.5 to 13.0 dB of gain (10.5 to 11.5 dB at 2.0 GHz), 4.5 to 5.0 dB noise figure and +5 to +6 dBm of output power at 1 dB gain compression. In 100 piece quantities, they are priced as low as $8.10 each for model MSA-0735, $23.55 each for model MSA-0770, and $2.45 each for model MSA-0785. **Avantek, Inc., Santa Clara, Calif. INFO/CARD #163.**

Network Analyzer Test Software: EEsof, Inc.

An advanced computer-aided test (CAT) software program for the calibration, measurement, and management of vector network analyzer data has been introduced. For use with both active and passive devices, EEsof's ANACAT™ is fully menu-driven and supports automatic network analyzers such as the Hewlett-Packard 8510, the HP 8753, and others. ANACAT's built-in database and calibration capabilities are enhanced by pop-up menus, mouse control, and interactive screen displays that assist the user in each step of the test process. The program has a DOS interface, full-screen editor for creating and revising files, and color graphics for interactive graphics display. In addition, the program's database readily interfaces with other popular applications such as EEsof's Touchstone®, Lotus' 1-2-3™, and Ashton-Tate's dBASE™.

ANACAT runs on the IBM PC-XT™, PC-AT™, Hewlett-Packard Vectra PC™, and compatibles. The program is priced between $5,000 and $7,500 depending upon options. **EEsof, Inc., Westlake Village, Calif. INFO/CARD #162.**

100 kHz-1 GHZ Signal Generator: Rohde & Schwart-Polarad, Inc.

The new signal generator SMX from Rohde & Schwartz-Polarad is a high qual-

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ity synthesizer, providing crystal referenced signals ranging from 100 kHz to 1000 MHz. The output level can be precisely adjusted from −137 dBm to +13 dBm in steps of 0.1 dB. The total level error of less than 1.5 dB down to −127 dBm allows accurate and reproducible sensitivity measurements on receivers. A non-interrupted RF output range of 10 dB, in 0.1 dB steps commencing at any RF level setting, is a standard feature of the SMX. Offering amplitude, frequency and pulse modulation, the SMX can be used for a great variety of measurements. AM and FM can be used separately, combined or as two tone modulation. Modulation depth and frequency deviation can be set independently of each other. **Rohde & Schwartz-Polarad, Inc., Lake Success, N.Y. INFO/CARD #161.**

GaAs Amplifier Chips: Harris Microwave Semiconductor

The first two gallium arsenide (GaAs) amplifier chips in an analog family are now available from Harris Microwave Semiconductor. These devices, the HMR-10502 and HMR-10503, give designers the option to add gain in the 500 MHz to 5 GHz range by using cascadable 10 dB gain blocks with ±0.75 dB full-band gain flatness. The HMR-10503 is a directly cascadable broadband amplifier chip that does not require external DC blocking on the RF input or output ports. The HMR-10502 is similar to the HMR-10503, but without internal source bypass circuitry; to let designers maximize flat gain in a particular band of interest by choosing the correct external source capacitor values. **Harris Microwave Semiconductors, Milpitas, Calif. INFO/CARD #153.**

Flush Mount 1 Watt Termination: EMC Technology, Inc.

A screw-in, flush mount 1 watt termination for microwave circuit packages has been introduced by EMC Technology. The new termination enhances package reliability by eliminating the problems caused by drilling accuracies and differential coefficient of expansion in press fit circuit packages. Model 4910 termination uses a standard SMA thread, and is ideal for use when testing SMA connectors in circuit packages. Frequency range is DC to 18 GHz, with 1 watt power at 60°C. VSWR is typically low, from 1.10 maximum at 4.0 GHz to 1.30 maximum at 18.0 GHz. Characteristic impedance is 50 ohms nominal. Price for the new Model 4910 is $9.00 (100s). **EMC Technology, Inc., Cherry Hill, N.J. INFO/CARD #152.**

Low-Cost Portable Oscilloscope: Tektronix, Inc.

Tektronix has broken the $1,000 price barrier with the introduction of the newest member of the popular 2200 Series of portable oscilloscopes. The dual channel 2225 Portable Oscilloscope offers features and performance including 50 MHz bandwidth, alternate magnification, 500 microvolt sensitivity, peak-to-peak auto trigger mode, and high frequency/low frequency trigger filtering. The scope's 500 microvolt sensitivity is four times more sensitive than previous 2200 scopes.

To provide improved low cost performance with the 2225, Tektronix offers the P6103, a 50 MHz, 10X passive probe for under $50. The P6103 has compensation built into the probe head and a more durable probe tip. The 2225 portable oscilloscope will be priced at $995. **Tektronix, Inc., Beaverton, Ore. Please circle INFO/CARD #158.**

Black Magic . . .

Ferrite—that black, crystalline oxide ceramic, offers the almost magical property of suppressing EMI—by itself—no grounds needed.

The lossy RF characteristics of ferrite absorb the EMI energy, converting it to harmless heat.

Slip a Fair-Rite suppressor element (shield bead) over a conductor, and EMI/RFI disappear. Yet DC doesn't know it's there.

In addition to the beads and sleeves which have become industry standards, Fair-Rite has developed a series of large 1 and 2 piece sleeves, large toroids, and shielding plates. All can be incorporated in new designs; many can be retrofitted to existing equipment. Data processing equipment, which both produces and is particularly vulnerable to EMI, offers a prime opportunity for the use of ferrites as EMI suppressors. Other applications include Switch Mode Power Supplies, telecommunication equipment, and designs that must meet TEMPEST requirements. A selection of these large suppressors is listed on the opposite page. Please contact Fair-Rite for other items in our broad line of EMI suppressor ferrites.

Fair-Rite, the leader in ferrite EMI suppressors, has developed unique capabilities in state-of-the-art shielding ferrites, offering users unparalleled experience in tooling development and the manufacturing process. To assist designers in making full use of our know-how, we offer three **ENGINEERING EVALUATION KITS:**

1. The **CABLE & CONNECTOR EMI SUPPRESSOR KIT** (part no. 0199000002): 13 different large beads and toroids, 1 and 2 piece sleeves, and the new multi hole shielding plates, supplied in Fair-Rite 43 material.
2. The **BEAD-ON-LEAD KIT** (part no. 0199000003): 10 different beads on leads, plus two six hole wound shield beads, providing impedances from 68 to 680 ohms typical, measured at 100 MHz. Values from 68 to 200 ohms are available taped and reeled for automatic insertion.
3. The **BEAD, BALUN AND BROADBAND KIT** (part no. 0199000001): A selection of 34 cores in 7 different materials offering the designer many options for high frequency circuits.

Each Kit is $25 postpaid against your check or M/O; or, plus shipping when ordered on your PO or letterhead. For prices outside the USA and Canada, please contact our Customer Service Department.

Each Kit comes with its own Engineering Bulletin describing the contents and guiding the designer in the use of Fair-Rite components to solve his EMI problems.

Further information on these three Kits may be obtained by requesting a copy of the Engineering Notes, *Ferrite Components as EMI Suppressors.*

For transformer and inductor designs: Fair-Rite manufactures a broad line of ferrite bobbins, U, E and I cores; pot, ETD, EP and PQ cores; toroids, slugs and rods. Send for our full line catalog.

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High Output Noise Generators: Noise Com Inc.

Noise Com has developed a new family of inexpensive high power RF White Noise Generators in a 24 PIN dual-in-line package. The high output level of these noise generators along with their excellent stability makes them ideal for injecting noise into systems with substantial decoupling for such applications as BITE-ERROR, amplifier gain and noise figure testing. Model NC2201 covers the full frequency range of 100 Hz-100 MHz with an output of +5 dBm and a flatness of ±1 dB, priced at $290 (1-9). Model NC2301 covers the frequency range of 1.5 MHz-300 MHz with an output of +0 dBm and a flatness of ±1 dB, and is priced at $330 (1-9). **Noise Com, Inc., Hackensack, N.J. INFO/CARD #157.**

Hybrid DIP VCXO: Vectron Laboratories, Inc.

The CO-484V Series Hybrid VCXOs (Voltage Controlled Crystal Oscillators) are designed specifically for phase locking applications. They provide +7 dBm output at any specified center frequency from 10 MHz through 200 MHz, and complement Vectron's previously announced 32 kHz to 70 MHz TTL output line. Deviation over 0 to +5V or ±5V control voltage is available from ±30 ppm to ±100 ppm; this deviation is adequate to permit locking onto the specified center frequency over the operating environment (0/50°C to −55/+85°C) for 10-20 years without adjustment. **Vectron Laboratories, Inc., Norwalk, Conn. INFO/CARD #160.**

Board-Mountable Enclosures: Compac Development Corp.

Compac introduces the FLT series of board-mountable RFI/EMI enclosures, in custom or standard configurations. The FLT enclosures are available with a one-cover option for direct access while installed. **Compac Development Corp., Holbrook, N.Y. INFO/CARD #159.**

High Power Mini Attenuators: Alan Industries, Inc.

Specifications for new high power additions to series 50 MHP attenuators include a 10 watt average power model with a frequency range of DC-18 GHz in a case size only 1.2 inches long and 0.615 inches in diameter. A 5-watt model is also available, with attenuation values of 1-20 dB. Both models are supplied with SMA connectors per MIL-C-39012. These fixed attenuators have a low VSWR of 1.35:1 maximum. Price for the 50 MHP series in lots of 1-24 is $65 each for the 5 watt model and $85 each for the 10 watt model. **Alan Industries, Inc., Columbus, Ind. INFO/CARD #156.**

MIL-Qualified Connectors: Applied Engineering Products

Sixty-five connectors are now qualified to MIL-C-39012 standards — 37 in SMA series, 17 in SMB series, 11 in SMC series. A 12-page brochure describes the available configurations, including cable plugs, bulkhead jacks and receptacles. Latest additions to the AEP connector line will be on display. **Applied Engineering Products, New Haven, Conn. Please circle INFO/CARD #155.**

Benchtop RF Pulse Amplifier: Amplifier Research

A new, very compact high-power RF amplifier has been developed especially for NMR and other applications that demand rapid blanking capability and instantly available bandwidth over the 2-150 MHz frequency band. The Model 1000LP amplifier delivers 1000 watts of linear pulse power on an up-to-10 percent duty cycle with a maximum eight-millisecond pulse. The fast blanking feature effectively shuts down RF power to permit sampling of NMR signals. In addition to pulse operation, the Model 1000LP has a front-panel CW pushbutton selector which may be used for continuous-wave applications at power levels to 200 watts. Pulse operation requires only a TTL gate pulse synchronized with the RF input signal. Price of the 1000LP is $13,500. **Amplifier Research, Souderton, Pa. Please circle INFO/CARD #154.**

Clip-on EMI Gasket: Chomerics, Inc.

Chomerics has introduced METAL-KLIP™ Clip-On EMI Gasketing products. This new clip-on method provides a simple means of mounting a carbon-loaded silicone or wire mesh EMI gasket strip on sheet metal enclosures. The corrosion-resistant stainless steel spring clip in which the gasket material is secured snaps onto the edge of a door or cabinet frame. Sharp teeth or tines inside the clip ensure electrical continuity between the METALKLIP assembly and the panel, and bite through non-conductive paint. The product is offered in standard 8-foot lengths or custom-cut. **Chomerics, Inc., Woburn, Mass. INFO/CARD #213.**

High Power Class A Amplifier: EPSCO, Inc.

The EPSCO Model AM0250D0015 is a miniature Class A Linear High Power Amplifier delivering 15 watts minimum at the 1 dB compression point across the full 20 to 500 MHz frequency band. **EPSCO, Inc., RF Division, Westlake Village, Calif. INFO/CARD #212.**

Push-Pull, TMOS Power FETs: Motorola Semiconductor

MRF151G (50V Vcc) and MRF141G (28V Vcc) are 300 watt, 175 MHz, TMOS power FETs in the GEMINI push-pull package. The lower voltage part has typical gain of 14 dB at 175 MHz, while the higher voltage part has 17 dB. **Motorola Semiconductor, Phoenix, Ariz. INFO/CARD #211.**

Surface Mount GaAs MESFETS: Motorola Semiconductor

MRFG9661,R and MRFG9801,R are low-noise dual gate GaAS MESFETS in 4-lead surface mount plastic packages. The MRFG9801 is designed for use at 500 MHz, while the MRFG9661 is optimized for best performance around 1 GHz. Noise figures are typically 1.2 dB; gains typically 17 dB. Both MRFG9661 and 9801 are SOT-143 packages. The MRFG9661,R and MRFG9801,R are in the conventional Japanese configuraton for 4-lead SOTs. **Motorola Semiconductor, Phoenix, Ariz. INFO/CARD #210.**

Pulse Power Amplifier: Steinbrecher Corp.

A low intermod pulse power amplifier operating over the 10 MHz to 86 MHz range, which provides a low-cost source of 400 watt PEP, 10 percent duty cycle pulses with intermod suppression of 28 to 30 dB across the entire frequency band, is available. **Steinbrecher Corp., Woburn, Mass. INFO/CARD #209.**

Low Noise Transistors: Motorola Semiconductor

Two series of state-of-the-art low-noise high frequency transistors: MRF940 Series (I_C = 15 mA) and MRF950 Series (I_C = 30 mA). Both series feature 1.2 micron interdigitated geometries, ion implantation, gold metalization and silicon nitride passivation. Package choices are the plastic macro-x (MRFXX1); the surface mount SOT-143 (MRFXXX1); a 100 mil. square metal/ceramic package (MRFXX2); and the surface mount SOT-23 (MMBRXX1). **Motorola Semiconductor, Phoenix, Ariz. INFO/CARD #208.**

NMR/MRI Kilowatt Amplifiers: Kalmus Engineering

Our new models 166LP (10-86 MHz) and 166 HP (80-200 MHz) are the world's only pulse kilowatt amplifiers capable of better than 1 microsecond blanking time, producing cleaner images. **Kalmus Engineering International, Ltd., Woodinville, Wash. INFO/CARD #207.**

Log and Compression Video Amplifiers: American Electronic Labs

Pulse-on-pulse and other custom hybrid log video amplifiers and compression video amplifiers with wide dynamic range and fast recovery. **American Electronic Laboratories, Inc., Lansdale, Pa. INFO/CARD #206.**

High Power RF Amplifiers: TIW Systems

VHP Series fixed gain high power RF amplifiers designed to meet high performance specifications. Spanning the 5 to 300 MHz frequency range, the new amplifiers feature greater than 30 dB gain, low VSWR in and out, and efficient thermal packaging which allows operation at elevated temperatures without additional heat sinking. Four models are available providing up to 1 watt linear output power, noise figures from 5 to 6 dB, and typical compression from +27 dBm to 30 dBm. Models VHP-01 and VHP-02 operate in the 40-300 MHz frequency range, while Models VHP-03 and VHP-04 cover the ranges of 5 to 250 MHz and 10 to 300 MHz, respectively. **TIW Systems, Inc., Sunnyvale, Calif. INFO/CARD #205.**

The Paramixer: Steinbrecher Corp.

The PARAMIXER™ supercomponent design provides a +82 dBm second order intercept and a +45 dBm third order intercept. Models currently in production provide this performance over a frequency range of 5 MHz to 30 MHz. The PARAMIXER provides the closest approximation to linear performance in frequency conversion. It therefore is useful as a key building block in designs involving frequency conversion where maximum dynamic range is required (e.g., receivers, translators for RF to digital conversion). **Steinbrecher Corp., Woburn, Mass. INFO/CARD #203.**

Synthesized Signal Generator: Rohde & Schwartz-Polarad

The SMG Synthesized Signal Generator covers 0.1 to 1000 MHz in 1 Hz steps, with an output level from −137 dBm to +13 dBm in steps of 0.1 dB. It offers exceptional spectral purity with residual FM of less than 1 Hz at 250 MHz and an SSB phase noise of −126 dBc at 20 kHz from the carrier at a frequency of 500 MHz. It offers high quality amplitude, frequency, phase, pulse and FSK modulation. An optional Modulation Synthesizer is available which can be used as an audio signal source from 10 Hz to 100 kHz. IEEE-488 interface is standard. **Rohde & Schwartz-Polarad, Lake Success, N.Y. INFO/CARD #202.**

Emission Measuring Receivers: Eaton Corp.

The new Eaton 3018 and 3038 Emission Measuring Receivers provide a cost effective means of making highly reliable FCC, VDE and CISPR measurements over the frequency range of 9 kHz to 30 MHz and 25 to 300 MHz, respectively. Designed for manual or programmable operation, the receivers feature full preselection, integral GPIB bus, and nonvolatile memory for storage and easy recall of front panel settings. **Eaton Corp., Electromagnetic Instrumentation Div., Los Angeles, Calif. INFO/CARD #201.**

Now, more than 135 dB spurious-free dynamic range in a broadband HF mixer.

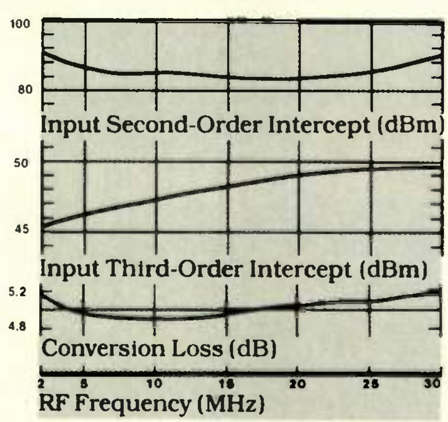
The patented STEINBRECHER PARAMIXER™ frequency converters have been designed for high performance HF communications systems. Tailored to either Receiver or Exciter applications, the PARAMIXER provides unequaled results for both requirements.

REDUCE OR ELIMINATE RECEIVER PRESELECTOR REQUIREMENTS

The RECEIVER PARAMIXER affords the best dynamic range currently available in a mixer. It also affords a low noise figure and full coverage of the HF band. The mixer can withstand continuous out-of-band interfering signals with peak power up to ONE WATT without desensitization.

REDUCE EXCITER BROADBAND NOISE AND POST-SELECTOR REQUIREMENTS

The companion EXCITER PARAMIXER maintains high dynamic range and can be operated continuously at input signal levels up to ONE WATT. Thus you can significantly reduce the power gain of your broadband amplifier and exciter broadband noise.



Typical PARAMIXER Parameters

The PARAMIXER concept incorporates a unique local oscillator power amplifier and wave-shaping circuit to achieve high dynamic range. The LO drive circuit, included in each PARAMIXER, requires only nominal drive power from your synthesizer.

PARAMIXER frequency converters are being used in modern communications systems, radar applications, and fast-hopping frequency translators.

Why not the BEST mixers for your HF system?

Typical PARAMIXER Specifications

Input Third-Order Intercept	> +45 dBm
Input Second-Order, Intercept	> +82 dBm
Noise Figure	< 6.5 dB
One-dB Compression Point	> +33 dBm
One-dB Suppression Point	> +33 dBm
Local Oscillator Power Required	< +20 dBm
DC Power Required	< 7 Watts

STEINBRECHER also offers complete digitally-controlled synthesizer and frequency translator designs compatible with the high performance of the PARAMIXER system.

STEINBRECHER engineers will custom design and build frequency translators with input bands ranging from 1 MHz to 50 GHz for your OEM applications.

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A Ladder Program Analysis Program

Personal Computers Both Small and Large Use This CAD Tool.

By Kenneth D. Wyatt
TRW Electronic Products, Inc.

Computer modeling of electrical networks is a real time-saver for RF engineers. Engineering CAD (Computer Aided Design) programs include basically two types of software: analytical and synthesis. Analytical software derives the circuit characteristics such as gain and phase from given component values. Synthesis software does just the reverse, starting with desired circuit characteristics such as filter type, bandwidth, and roll-off to derive the component values required. The program included with this article analyzes a passive ladder or cascaded network. One valuable use of this program is analysis and verification of existing circuit designs. Using the editor function, it is also possible to "tweak" component values in order to optimize a circuit design.

Most RF networks can be modeled as a series of cascaded circuit elements such as series or parallel (shunt) resistors, inductors or capacitors. The 17 circuit models included in this program may be cascaded to form almost any type of filter, impedance matching network, or transmission line network.

The program allows you to enter component values and obtain a chart of insertion loss, return loss, VSWR, reflection coefficient (Rho), and both real and imaginary input impedances. Circuits may be completely modeled before the breadboard stage, saving much time with hand calculations or trial and error in the lab.

The included program is designed to run on the IBM-PC or compatibles under BASICA or GWBASIC. Except for the plotting routine, special graphics or coding was avoided so that it might be easily converted for other computers that use Microsoft BASIC. In addition, a version of the program is available for the Commodore 64 or 128 computers.

Theory of Operation

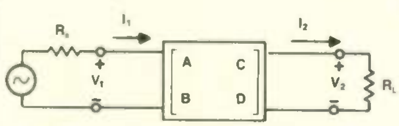
The program is based upon the ABCD parameters of the circuit element to be analyzed. This is a standard network analysis technique, but one that may not

Type 1	Series R	$\begin{bmatrix} 1 + j0 & R + j0 \\ 0 + j0 & 1 + j0 \end{bmatrix}$
Type 2	Parallel R	$\begin{bmatrix} 1 + j0 & 0 + j0 \\ \frac{1}{R} + j0 & 1 + j0 \end{bmatrix}$
Type 3	Series L	$\begin{bmatrix} 1 + j0 & 0 + j(\omega L) \\ 0 + j0 & 1 + j0 \end{bmatrix}$
Type 4	Parallel L	$\begin{bmatrix} 1 + j0 & 0 + j0 \\ 0 - j\left(\frac{1}{\omega L}\right) & 1 + j0 \end{bmatrix}$
Type 5	Series C	$\begin{bmatrix} 1 + j0 & 0 - j\left(\frac{1}{\omega C}\right) \\ 0 + j0 & 1 + j0 \end{bmatrix}$
Type 6	Parallel C	$\begin{bmatrix} 1 + j0 & 0 + j0 \\ 0 + j(\omega C) & 1 + j0 \end{bmatrix}$
Type 7	Series — Series RLC	$\begin{bmatrix} 1 + j0 & R + j\left(\frac{\omega L - 1}{\omega C}\right) \\ 0 + j0 & 1 + j0 \end{bmatrix}$
Type 8	Parallel — Series RLC	$\begin{bmatrix} 1 + j0 & 0 + j0 \\ \left[\frac{\omega^2 C^2 R}{\omega^2 C^2 R^2 + (\omega^2 LC - 1)^2}\right] - j\,(IC) & 1 + j0 \end{bmatrix}$ where: $IC = \frac{\omega C\,(\omega^2 LC - 1)}{\omega^2 C^2 R^2 + (\omega^2 LC - 1)^2}$
Type 9	Series — Parallel RLC	$\begin{bmatrix} 1 + j0 & RB + j(IB) \\ 0 + j0 & 1 + j0 \end{bmatrix}$ where: $RB = \frac{\omega^2 RL^2}{\omega^2 R^2 L\,(1 - \omega^2 LC)^2}$, $IB = \frac{\omega LR^2\,(1 - \omega^2 LC)}{\omega^2 L^2 R^2\,(1 - \omega^2 LC)^2}$
Type 10	Parallel — Parallel RLC	$\begin{bmatrix} 1 + j0 & 0 + j0 \\ \frac{1}{R} - j\left(\frac{1 + \omega^2 LC}{\omega L}\right) & 1 + j0 \end{bmatrix}$
Type 11	Series — Series RL/Series — Parallel C	$\begin{bmatrix} 1 + j0 & RB + j(IB) \\ 0 + j0 & 1 + j0 \end{bmatrix}$ where: $RB = \frac{R}{(1 - \omega^2 LC)^2 + (\omega RC)^2}$, $IB = \frac{\omega\,[L\,(1 - \omega^2 LC) - CR^2]\,RB}{R}$
Type 12	Transmission Line	$\begin{bmatrix} \cos(\theta) + j0 & 0 + j\,[Z_o \sin(\theta)] \\ 0 + j\left[\frac{\sin(\theta)}{Z_o}\right] & \cos(\theta) + j0 \end{bmatrix}$ where: $\theta = \frac{\sqrt{\omega L \epsilon_r}}{3E10}$
Type 13	Open Parallel Stub	$\begin{bmatrix} 1 + j0 & 0 + j0 \\ 0 + j\left[\frac{\tan(\theta)}{Z_o}\right] & 1 + j0 \end{bmatrix}$ where: $\theta = \frac{\sqrt{\omega L \epsilon_r}}{3E10}$
Type 14	Shorted Parallel Stub	$\begin{bmatrix} 1 + j0 & 0 + j0 \\ 0 - j\left[\frac{1}{\tan(\theta)\,Z_o}\right] & 1 + j0 \end{bmatrix}$ where: $\theta = \frac{\sqrt{\omega L \epsilon_r}}{3E10}$
Type 15	Open Series Stub	$\begin{bmatrix} 1 + j0 & 0 - j\left[\frac{Z_o}{\tan(\theta)}\right] \\ 0 + j0 & 1 + j0 \end{bmatrix}$ where: $\theta = \frac{\sqrt{\omega L \epsilon_r}}{3E10}$
Type 16	Shorted Series Stub	$\begin{bmatrix} 1 + j0 & 0 + j\left[\frac{Z_o}{\tan(\theta)}\right] \\ 0 + j0 & 1 + j0 \end{bmatrix}$ where: $\theta = \frac{\sqrt{\omega L \epsilon_r}}{3E10}$
Type 17	Transformer n:1	$\begin{bmatrix} n + j0 & 0 + j0 \\ 0 + j0 & \frac{1}{n} + j0 \end{bmatrix}$ where: n = turns ratio

Table 1 — Circuit Models

have been explained adequately in previous literature. The advantage of the ABCD parameters lies in the ease in which cascaded networks may be represented and analyzed.

The ABCD parameters make up a matrix that describes the voltages and currents into and out of four terminal (two port) networks (Figure 1). Each element model has a unique ABCD matrix as shown in Table 1. The operation of this program is based on the fact that the ABCD matrix of two cascaded circuits is equal to the product of their individual ABCD matrices. See Figure 2 for a brief derivation of this technique. These matrices are stored as the various element models, and their associated component values are entered by the user. At each frequency to be analyzed, the individual matrices are formed and multiplied to gradually compute the overall matrix of the entire circuit. Once the network is reduced to a single matrix, we may derive the insertion loss, return loss, VSWR, reflection coefficient, and input impedance. Table 2 illustrates the equations used in the program.



$V_1 = A\,V_2 + B\,I_2$
$I_1 = C\,V_2 + D\,I_2$
or, in matrix form:

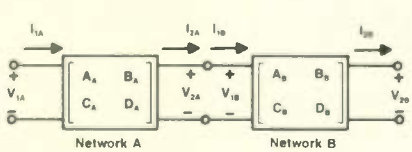
$$\begin{bmatrix} V_1 \\ I_1 \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix} \begin{bmatrix} V_2 \\ I_2 \end{bmatrix}$$

Figure 1. Two-port ABCD matrix.

Program Description

The program is divided into four sections: data input, data editor, computation and data output. Since the screen on the Commodore 64 is only 40 columns wide, this version displays just four of the most widely used parameters on the screen (frequency, gain, return loss and VSWR). These variables, plus reflection coefficient and input impedance, are all sent to the printer port if the print function is selected. In addition, the gain and return loss may be plotted graphically to the screen. Since the IBM-PC does not have this screen width limitation, the PC version of this program sends all variables to either the screen or printer.

Units used have been optimized for use at RF and are in ohms, nanohenries, picofarads, inches, degrees and megahertz. The maximum number of circuit sections has been limited to 30, but this is really arbitrary as most computers have plenty of memory these days. For passive net-



Note that $I_{1B} = I_{2A}$ and $V_{1B} = V_{2A}$

then: $\begin{bmatrix} V_{1A} \\ I_{1A} \end{bmatrix} = \begin{bmatrix} A_A & B_A \\ C_A & D_A \end{bmatrix} \begin{bmatrix} V_{2A} \\ I_{2A} \end{bmatrix}$

since: $\begin{bmatrix} V_{2A} \\ I_{2A} \end{bmatrix} = \begin{bmatrix} V_{1B} \\ I_{1B} \end{bmatrix} = \begin{bmatrix} A_B & B_B \\ C_B & D_B \end{bmatrix} \begin{bmatrix} V_{2B} \\ I_{2B} \end{bmatrix}$

then: $\begin{bmatrix} V_{1A} \\ I_{1A} \end{bmatrix} = \begin{bmatrix} A_A & B_A \\ C_A & D_A \end{bmatrix} \begin{bmatrix} A_B & B_B \\ C_B & D_B \end{bmatrix} \begin{bmatrix} V_{2B} \\ I_{2B} \end{bmatrix}$

The generalization to n cascaded 2-port networks follows:

$$\begin{bmatrix} A & B \\ C & D \end{bmatrix} = \begin{bmatrix} A & B \\ C & D \end{bmatrix}_1 \begin{bmatrix} A & B \\ C & D \end{bmatrix}_2 \cdots \begin{bmatrix} A & B \\ C & D \end{bmatrix}_n$$

Figure 2. Derivation of *n* cascaded networks.

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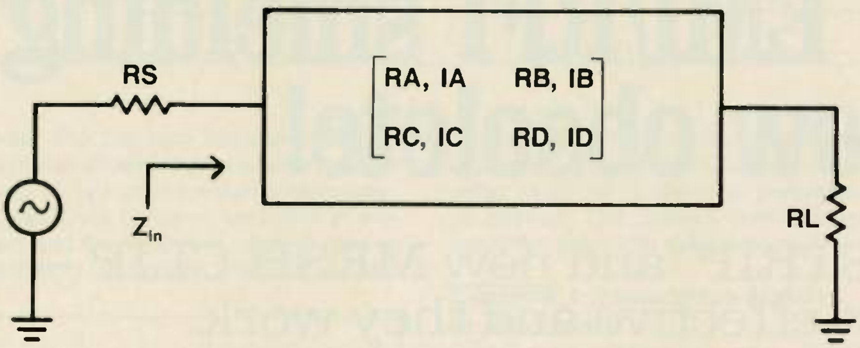
Given the result matrix for the network:



where: RA = the real part of element A (same for B, C, D)
IA = the imaginary part of element A (same for B, C, D)
RS = the source resistance
RL = the load resistance
Z_{in} = the input impedance

then let:

$$G = \frac{RA\,(RL) + RB + RC\,(RS)(RL) + RD\,(RS)}{(RL)}$$

$$H = \frac{IA\,(RL) + IB + IC\,(RS)(RL) + ID\,(RS)}{(RL)}$$

$$I = \sqrt{G^2 + H^2}$$

$$V1 = RC\,(RL) + RD$$

$$V2 = RC\,(RL) + ID$$

$$Q = V1^2 + V2^2$$

then we can now calculate:

$$\text{Insertion Loss (IL)} = -10 \log \left[\frac{(RL)\ I}{4\ (RS)}\right]$$

$$\text{Real part of } Z_{in} = \frac{V1\,[RA(RL) + RB] + V2\,[IA(RL) + IB]}{Q} = R_e\,[Z_{in}]$$

$$\text{Imaginary part of } Z_{in} = \frac{V1\,[IA(RL) + IB] + V2\,[RA(RL) + RB]}{Q} = I_m\,[Z_{in}]$$

$$\text{Rho } (\varrho) = \frac{\sqrt{\{R_e\,[Z_{in}] - RS\}^2 + I_m\,[Z_{in}]^2}}{\sqrt{\{R_e\,[Z_{in}] + RS\}^2 + I_m\,[Z_{in}]^2}}$$

$$\text{Return Loss (RL)} = 20 \log (1/\varrho)$$

$$\text{VSWR} = \frac{1 + \varrho}{1 - \varrho}$$

Table 2. Equations Used

work analysis, the insertion loss is equal to the transducer power gain. Thus, when the source (RS) and load (RL) resistances are matched, the gain is zero dB. ZIN(R) and ZIN(I) are the real and imaginary input impedances, respectively. The remaining output variables are RL (return loss in dB), VSWR (voltage standing wave ratio), and RHO (reflection coefficient).

Before the program is run, it is useful to prepare the network for analysis in order to ease data entry. Refer to Example 1 (Figure 3(a)), and notice that the circuit is drawn such that all elements are in cascade or "inline." The source resistance (RS) should always be drawn in

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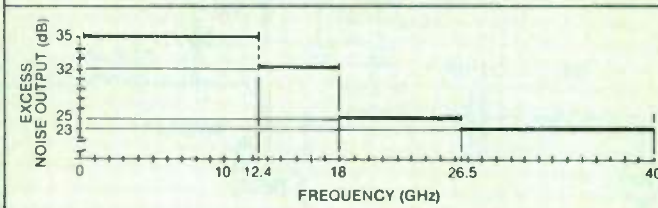

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MC 1100	10 - 1500	15.5dB	± 0.50dB	+ 28V, 10mA
MC 1040/1040P	10 - 4000	25.5dB/15.5dB	± 0.50dB	+ 28V, 15mA
MC 5112/5112P	1000 - 12400	25.5dB/15.5dB	± 0.50dB	+ 28V, 15mA
MC 5118/5118P	1000 - 18000	25.5dB/15.5dB	± 0.50dB	+ 28V, 15mA
MC 50018/50018P	10 - 18000	25.5dB/15.5dB	± 0.75dB	+ 28V, 15mA

WAVE GUIDE BAND SOURCES (GAS TUBE REPLACEMENTS)				
MC 51218W	12400 - 18000	15.0dB	± 0.50dB	+ 28V, 15mA
MC 51826W	18000 - 26500	25.0dB	± 2.00dB	+ 28V, 20mA
MC 52640W	26500 - 40000	23.0dB	± 3.00dB	+ 28V, 20mA

HERMETIC NOISE SOURCES				
MC 7090H	8325 - 9675	35.0dB	± 0.50dB	+ 28V, 15mA
MC 7146H	13500 - 15700	30.0dB	± 0.50dB	+ 28V, 15mA

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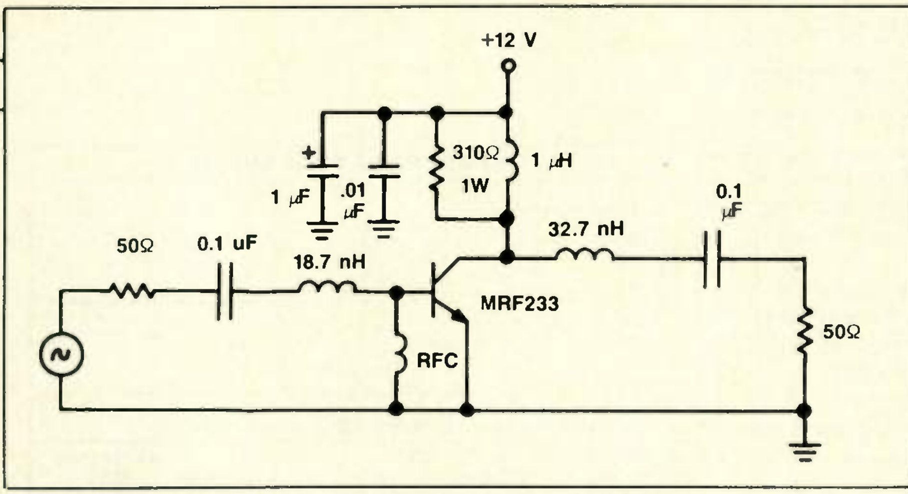

Figure 4(a). Example 2: Matching to a complex load.

(3) RHO, or (4) real/imaginary impedances. Up to 150 points may be plotted vs. frequency. The C64 version is limited to 40 points of insertion and return loss. You only need to specify the minimum and maximum desired for the vertical (Y) scale. You may also redefine the frequency limits, resweep the network, and replot the data. If the IBM program "GRAPHICS.COM" has been initially loaded into memory, the output plots of the IBM-PC version may be printed on any Epson graphics printer. Simply press the "SHIFT" and "PrtSc" keys simultaneously after the plot is complete on the screen. Press any key (IBM-PC version) to obtain the menu of choices after the plot is finished.

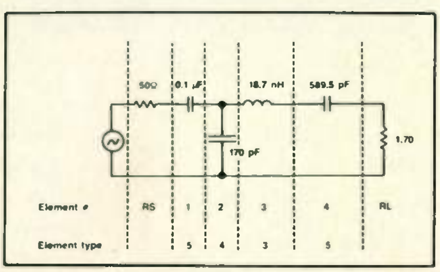

Figure 4(b). Example 2: input model.

In the case of a data entry error, you may edit the source or load resistances or any of the network components. Choice 5 allows you to exit from the program.

Example 2: Impedance Match to a Complex Load

Let us analyze the input and output matching circuits shown in the amplifier circuit of Figure 4(a). The source and load resistances are 50 ohms and the transistor used is a Motorola MRF233. First, we will verify the insertion loss, the input return loss, and input VSWR and the 3 dB bandwidth of the input circuit at 100 MHz. The series input impedance model for this device is 1.7-j2.7 ohms at 100 MHz. This transforms to a series equivalent of 1.7 ohms and 589.5 pF.

First, redraw the input model as shown in Figure 4(b). Prepare the model for data input by dividing the element types and recording the component values. In this case, show the transistor input capacitance (589.5 pF) as a separate element. Then enter all data as before. Sweep the circuit from 50 to 150 MHz in steps of 5 MHz. This will require a total of 21 steps.

Use either the (S)creen or (P)rinter option as desired. The results are shown in Figure 4(c). You will note that the circuit is matched at 100 MHz. The input return loss is 52.14 and the input VSWR is 1.00:1. The 3 dB bandwidth is about 30 MHz.

Similarly, the output impedance matching circuit may be analyzed for proper match at 100 MHz, output return loss, and output VSWR. The output impedance of the MFR233 may be represented by a series connected 5 ohm resistance and a 284.2 pF capacitance. Again model the transistor output capacitance as a separate circuit element. Refer to Figure 5(a) for the resulting output model. The results may be seen in Figure 5(b) and plotted in Figure 5(c). Note that the RF choke and +12 V feed need not be modeled, having little contribution to the system impedance.

Example 4: Transmission Line Match

This next example is an input impe-

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NETWORK LISTING:

ELEM        TYPE        R          L          C
                       (ZO)       (L)        (E)

SOURCE R                50
 1           5          0          0       100000
 2           6          0          0          170
 3           3          0        18.7           0
 4           5          0          0        589.5
LOAD R                  1.7

NOTE:   R,L,C ARE FOR ELEMENT TYPES 1-11.
        ZO,L,E ARE FOR ELEMENT TYPES 12-16.

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```

FREQ(MHZ)	IL(DB)	RL(DB)	VSWR	RHO	ZIN(R)	ZIN(I)
50.00	-8.78	-0.62	28.25	0.93	1.77	0.29
55.00	-8.24	-0.71	24.65	0.92	2.03	1.46
60.00	-7.60	-0.83	20.98	0.91	2.39	2.71
65.00	-6.88	-1.00	17.42	0.89	2.89	4.12
70.00	-6.04	-1.24	14.00	0.87	3.62	5.78
75.00	-5.08	-1.62	10.79	0.83	4.75	7.82
80.00	-3.98	-2.22	7.87	0.77	6.63	10.41
85.00	-2.77	-3.27	5.38	0.69	10.03	13.77
90.00	-1.51	-5.31	3.37	0.54	16.90	17.69
95.00	-0.45	-10.07	1.91	0.31	31.48	18.54
100.00	0.00	-52.14	1.00	0.00	49.87	-0.21
105.00	-0.50	-9.60	1.99	0.33	37.42	-27.63
110.00	-1.80	-4.69	3.79	0.58	18.30	-29.62
115.00	-3.44	-2.61	6.69	0.74	9.36	-24.81
120.00	-5.12	-1.60	10.91	0.83	5.37	-20.60
125.00	-6.70	-1.04	16.66	0.89	3.37	-17.50
130.00	-8.15	-0.72	24.07	0.92	2.27	-15.22
135.00	-9.47	-0.52	33.32	0.94	1.61	-13.49
140.00	-10.68	-0.39	44.88	0.96	1.18	-12.15
145.00	-11.79	-0.30	58.28	0.97	0.90	-11.07
150.00	-12.82	-0.23	74.40	0.97	0.70	-10.19

Figure 4(c). Example 2 input model data.

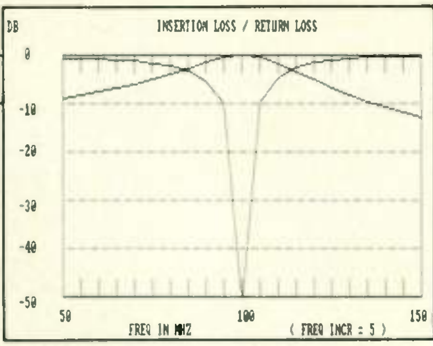

Figure 4(d). Example 2 input model response plot.

dance match to a Motorola MRF627 using a transmission line. See Figure 6(a). The equivalent input impedance for this device at 470 MHz is 6.0-j4.0 ohms. This equates to a series RC circuit of 6 ohms and 84.6 pF. The matching circuit uses a series transmission line and two variable capacitors which tune from 1 to 25 pF, each. The transmission line is 0.25 inches wide by 1.75 inches long. The circuit board is Teflon-fiberglass which has a relative dielectric constant of 2.55 and a thickness of 0.06 inches.

There are two input options for stubs and transmission lines. In some cases the actual physical dimensions are specified or may be measured from the circuit board, and in some cases the electrical parameters are specified in ohms and degrees of length (360 degrees = one wavelength). You will be given the option as to which set of data you desire. In both cases, the relative dielectric constant of the board material must be specified. The program will automatically calculate the effective dielectric constant. The effective dielectric constant differs from the relative dielectric constant depending upon the line impedance and board thickness.

In this case, choose (1) Physical Dimensions. Next, enter a dielectric constant of 2.55, transmission line dimensions of 0.25 and 1.75 inches, and board thickness from 400 to 540 MHz in steps of 10 MHz (16 steps). You should obtain the result shown in Figure 6(c). Note that the insertion loss is very broad band and varies from zero to 1.1 dB. The return loss at the design center frequency is 37.52 dB, and the VSWR at band edges is less than 2.79:1. By "tuning" the two variable capacitors with the edit function, you may observe the corresponding changes in circuit parameters, such as insertion loss and return loss.

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RF NETWORK ANALYSIS PROGRAM - VER 4.0PC
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NETWORK LISTING:

ELEM	TYPE	R (Z0)	L (L)	C (E)
SOURCE R		50		
1	5	0	0	100000
2	6	0	0	95.3
3	3	0	32.7	0
4	5	0	0	284.2
LOAD R		5		

NOTE: R,L,C ARE FOR ELEMENT TYPES 1-11.
Z0,L,E ARE FOR ELEMENT TYPES 12-16.

```
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```

FREQ(MHZ)	IL(DB)	RL(DB)	VSWR	RHO	ZIN(R)	ZIN(I)
50.00	-5.08	-1.62	10.79	0.83	4.64	-1.61
55.00	-4.64	-1.83	9.54	0.81	5.24	0.24
60.00	-4.16	-2.10	8.31	0.79	6.03	2.11
65.00	-3.63	-2.47	7.08	0.75	7.11	4.09
70.00	-3.05	-2.97	5.90	0.71	8.61	6.23
75.00	-2.43	-3.68	4.79	0.65	10.77	8.57
80.00	-1.78	-4.73	3.77	0.58	13.98	11.09
85.00	-1.15	-6.33	2.86	0.48	18.93	13.55
90.00	-0.58	-9.00	2.10	0.35	26.68	14.98
95.00	-0.17	-14.22	1.48	0.19	38.02	12.48
100.00	0.00	-42.10	1.02	0.01	49.67	0.71
105.00	-0.16	-14.48	1.47	0.19	50.37	-19.29
110.00	-0.65	-8.57	2.19	0.37	37.89	-32.83
115.00	-1.42	-5.55	3.24	0.53	24.55	-35.34
120.00	-2.36	-3.77	4.68	0.65	15.65	-32.94
125.00	-3.40	-2.66	6.59	0.74	10.32	-29.56
130.00	-4.45	-1.93	9.05	0.80	7.09	-26.42
135.00	-5.50	-1.44	12.11	0.85	5.07	-23.76
140.00	-6.51	-1.10	15.83	0.88	3.75	-21.57
145.00	-7.47	-0.86	20.29	0.91	2.85	-19.74
150.00	-8.39	-0.68	25.52	0.92	2.22	-18.22

Figure 5(b). Example 2 output model data.

Program Notes

The program described has proven to be a very useful tool in analyzing existing circuit designs. A number of benchmark networks have been tested and compared with other network analysis programs with gratifying results. Every effort has been made to eliminate errors, however, if the reader should discover a "bug," please write the author at the address below. Enclose a self-addressed stamped envelope if you desire a reply.

It is possible to change the default units by changing the multiplication factors in

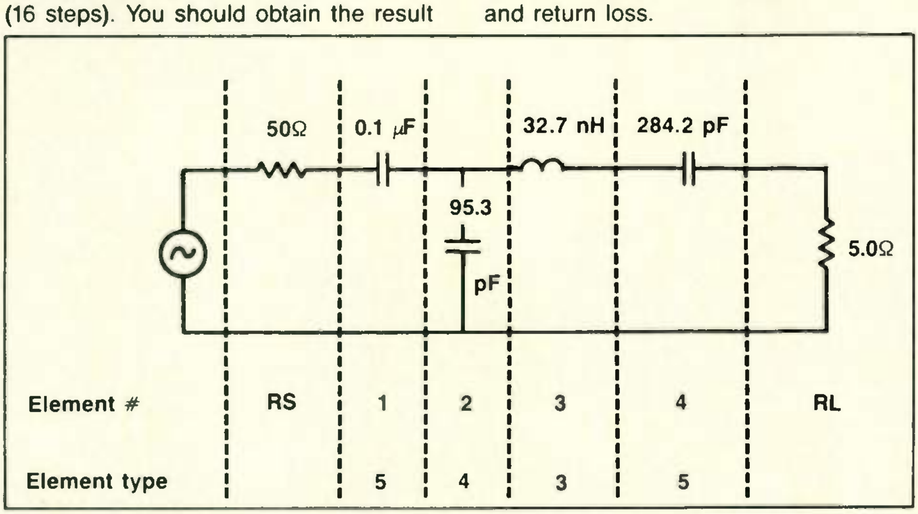


Figure 5(a). Example 2: output circuit.

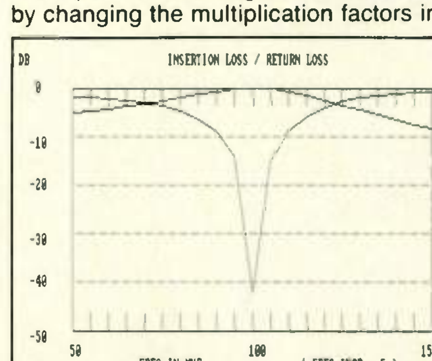
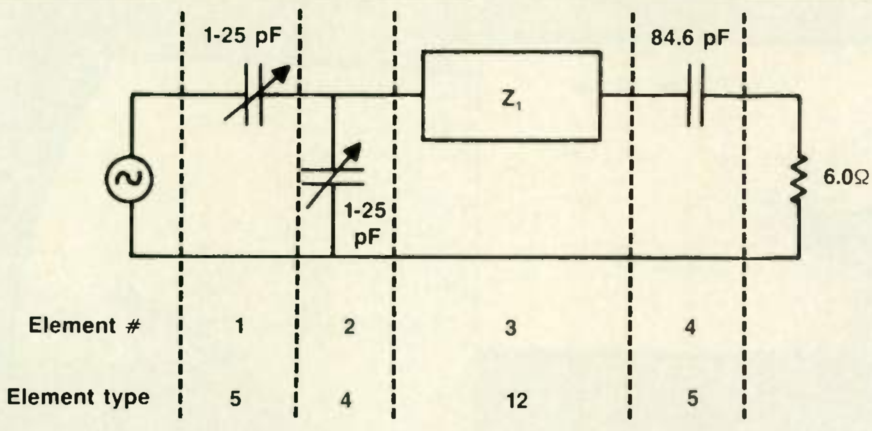


Figure 5(c). Example 2 output model response plot.



Notes: 1) Z_1 = 0.25" wide × 1.75" long
2) The circuit board is Teflon-Fiberglass and has a relative dielectric constant of 2.55 and the thickness is 0.06".
3) Both variable capacitors were set to 9 pF for the best match.

Figure 6(a). Example 4: Transmission line match.

Line 320 (310 for C64 version). Note, however, that if the frequency is changed to something other than MHz, you will have to change the various "MHz" labels within the program.

Due to formatting constraints, input data larger than 6 digits to the left of the decimal is not allowed and will default to 999999.00. Similarly, output data larger than 6 digits to the left of the decimal point will have just the leftmost 8 digits displayed.

Input data of zero defaults to 0.00001 (1E-5) in order to avoid divide by zero errors in the equations.

When typing in the program, watch out for the following variable names: CO, RO, ZO and OUT$. They contain the letter O

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RF NETWORK ANALYSIS PROGRAM - VER 4.0PC
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NETWORK LISTING:

ELEM        TYPE          R           L           C
                        (ZO)         (L)         (E)

SOURCE R                  50
 1            5            0           0           9
 2            6            0           0           9
 3           12        38.15        1.75        2.55
 4            5            0           0    84.66001
LOAD R                     6

NOTE:    R,L,C ARE FOR ELEMENT TYPES 1-11.
         ZO,L,E ARE FOR ELEMENT TYPES 12-16.

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 FREQ(MHZ)   IL(DB)   RL(DB)    VSWR     RHO    ZIN(R)   ZIN(I)

   400.00    -1.10    -6.52     2.79    0.47    20.53   -17.59
   410.00    -0.82    -7.66     2.41    0.41    22.89   -14.57
   420.00    -0.57    -9.08     2.08    0.35    25.69   -11.54
   430.00    -0.37   -10.90     1.80    0.28    29.02    -8.53
   440.00    -0.21   -13.36     1.55    0.21    33.02    -5.60
   450.00    -0.09   -16.95     1.33    0.14    37.84    -2.86
   460.00    -0.02   -23.32     1.15    0.07    43.63    -0.49
   470.00     0.00   -37.52     1.03    0.01    50.57     1.21
   480.00    -0.03   -21.69     1.18    0.08    58.77     1.79
   490.00    -0.10   -16.24     1.36    0.15    68.21     0.62
   500.00    -0.22   -13.02     1.58    0.22    78.55    -3.11
   510.00    -0.38   -10.78     1.81    0.29    88.93   -10.25
   520.00    -0.57    -9.11     2.08    0.35    97.84   -21.25
   530.00    -0.79    -7.80     2.38    0.41   103.32   -35.65
   540.00    -1.03    -6.75     2.70    0.46   103.79   -51.70
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Figure 6(b). Example 4 data.

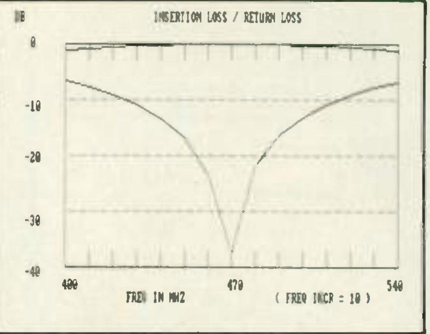

Figure 6(c). Example 4 response plot.

and not the number zero (0).

The author will provide a copy of the program on disk for a cost of $10.00, including disk and postage. Requests should be sent to Ken Wyatt, 56 Aspen Drive, Woodland Park, CO 80863. Readers wishing only a copy of the program listing should send a self-addressed, stamped business size envelope to *RF Design*, 6530 South Yosemite St., Englewood, CO 80111. Specify either the IBM or C-64 version.

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About the Author

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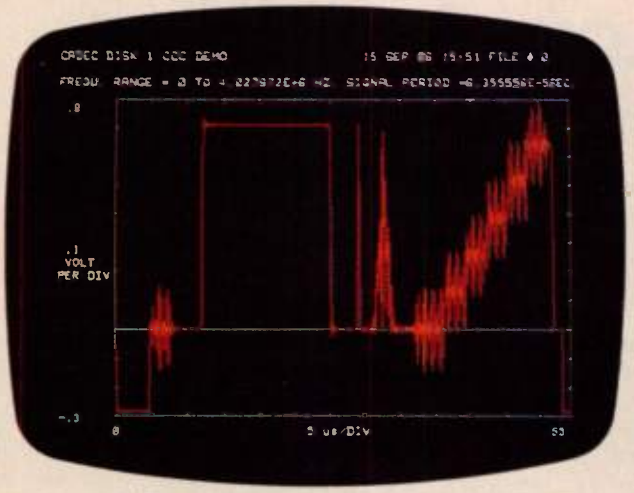

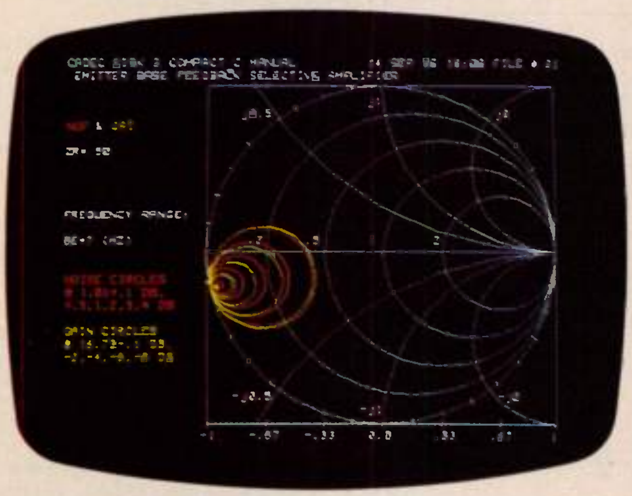

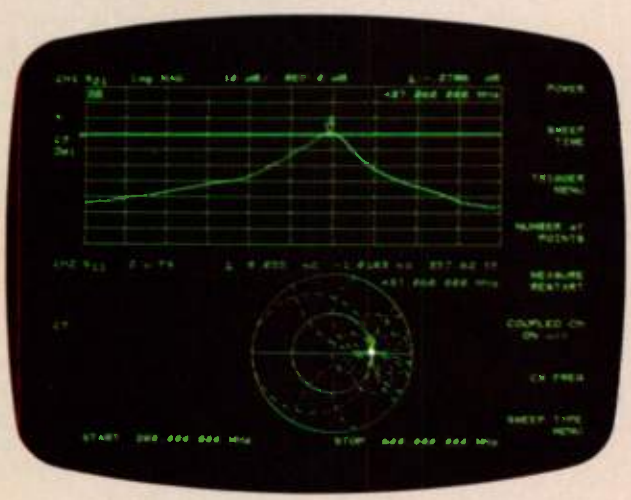

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INFO/CARD 66

An Eight-Stage Log Amplifier

A Review of Log Techniques and the Design of a Successive-Detection System

By Diep Tran-Nguyen
Plessey Semiconductors

In the demanding world of radar, sonar and ECM, the high pulse densities, narrow pulse widths and rapid amplitude variations cannot be dealt with by linear, limiting, AGC amplifier systems. The only systems that can deal with these signals utilize logarithmic amplifiers. Those amplifiers, often called log amps, are not only capable of handling compression of high input dynamic range and instantaneous pulse to pulse response, but also the required high sensitivity. They amplify signals logarithmically, with output voltage proportional to the logarithm of the input. In a typical log amplifier system, an input dynamic range of 80 dB can be compressed to about 20 dB.

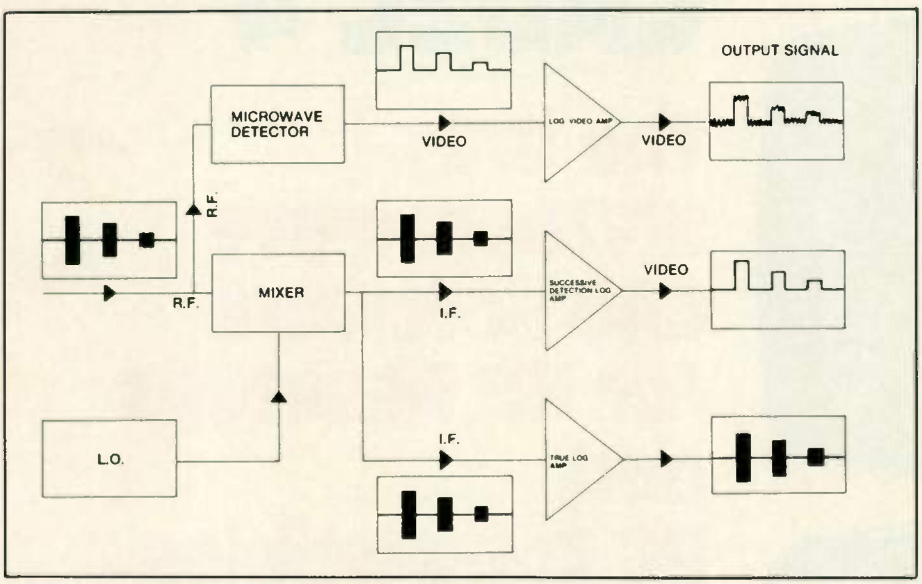

Figure 1. Pulse response of three-log amplifiers types.

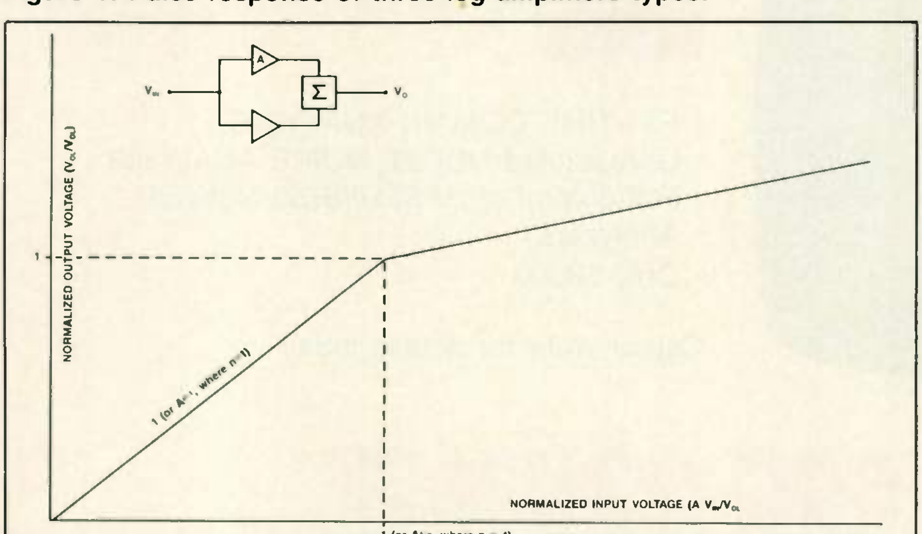

Figure 2. Transfer function of a true log amplifier

There are three types of logarithmic amplifiers: the *detector log video amplifier* (or log video amplifier), the *true log amplifier,* and the *successive detection log amplifier.* To distinguish between the amplification characteristics of these different log systems, refer to Figure 1, which shows the output signals of each log receiver system, with an RF pulse input.

Although this article presents a design example of a successive log system, design details will also be discussed for a true log system and a log video system.

Log Video Systems

In a log video system, the amplifiers perform a log function directly on the video signal from a microwave detector, operating within the bandwidth of the demodulated signal. These log video systems can provide broad coverage through microwave since the frequency response is determined only by the detector. However, this system has several disadvantages. Sensitivity is limited to about −45 dBm by the detector circuit and although it is possible to get a log video system with a dynamic range of 80 dB, the overall dynamic range is reduced to half by the square law of the detector. Coupling between the video amplifiers can lead to a potential problem: If they are AC coupled they won't respond to a CW signal, and if they are DC coupled they become prone to DC drift due to the high DC gain requirement.

A recent log video amplifier design described by Potson and Hughes (1) uses Schottky diode bridges as non linear elements and state-of-the-art op amps in which bandwidth, rise and fall times are independent of the closed loop gain. This technique employs the bridge small signal resistance and limit current to realize the log response.

Log video amplifiers are often found in direction finding and alarm systems rather than in more sensitive measurement equipment.

True Log Systems

True log amplifiers perform a logarithmic function at IF and have undetected, undistorted IF outputs. They are used in applications where both sensitivity and low distortion in the IF are required. A dual gain technique is used to obtain the log transfer function. Each amplifier stage consists of a limiting amplifier with a gain A = 3.16 (10 dB) which is in parallel with a unity gain amplifier. At low signal levels, the gain is linear at 10 dB. When the input signal exceeds a certain level the gain is then dropped to unity. As shown in Figure 2, the output V_0 is normalized to V_{ol} and the input is normalized to V_{ol}/A, where V_{ol} is output limited level.

Figure 3 shows the transfer function of a six-stage true log system. The transfer function consists of a series of straight lines with break points where each limiting amplifier successively reduces the gain by a factor of A. Note that the slopes of the straight line and break points on the horizontal axis are derived from the definition of the dual gain stage. The multiple scales of the (1-1/A) on the vertical axis are found by setting the two equations of the two straight lines equal at the intersection points.

Figure 4 shows a log transfer characteristic of a true log amplifier which consists of n stages, each with a low level gain stage A = 4 (12 dB). The log nonlinearity referred to the input in dB can be determined by drawing the two straight lines, one connecting the cusps of the cascade's characteristic and the other connecting tangentially to the characteristic between cusps. For a small signal gain of A = 3.16 (10 dB), the log nonlinearity is 1.4 dB.

A typical 6 stage-true log amplifier "strip" using Plessey SL531C true log amplifiers can give 70 dB of a dynamic range from low frequencies below 1 kHz (using external decoupling capacitors) to higher frequencies up to 200 MHz, and a typical phase shift of 3 degrees (2).

Since the stages are operating at RF and no detection takes place, this type of log amp has the advantage that the carrier information is preserved. The disadvantage is that its operating frequency is relatively low. With their low phase shift and low frequency operation, these log

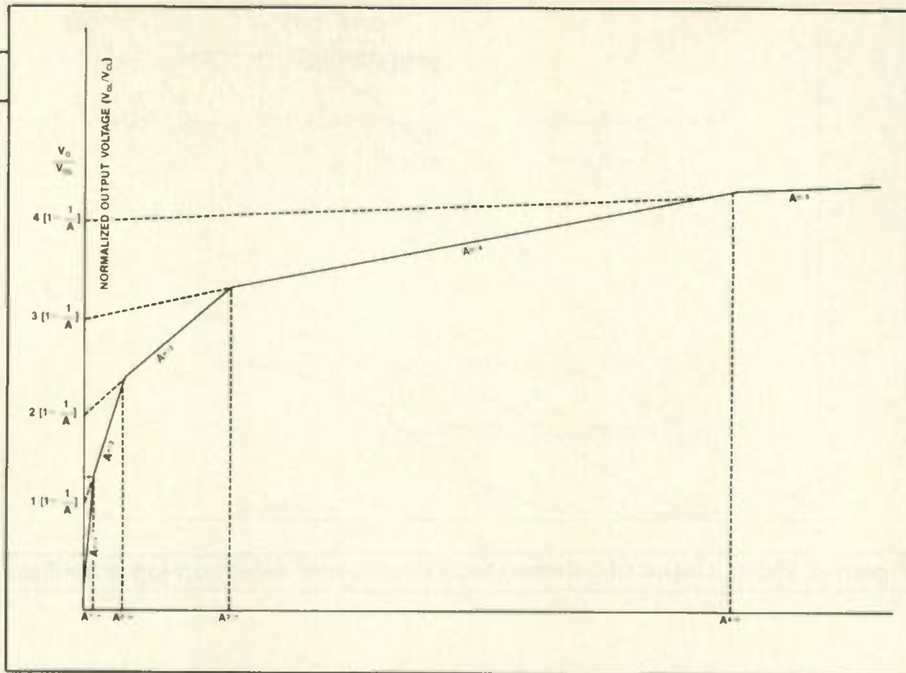

Figure 3. Transfer function of an N-stage true log IF strip (linear scale).

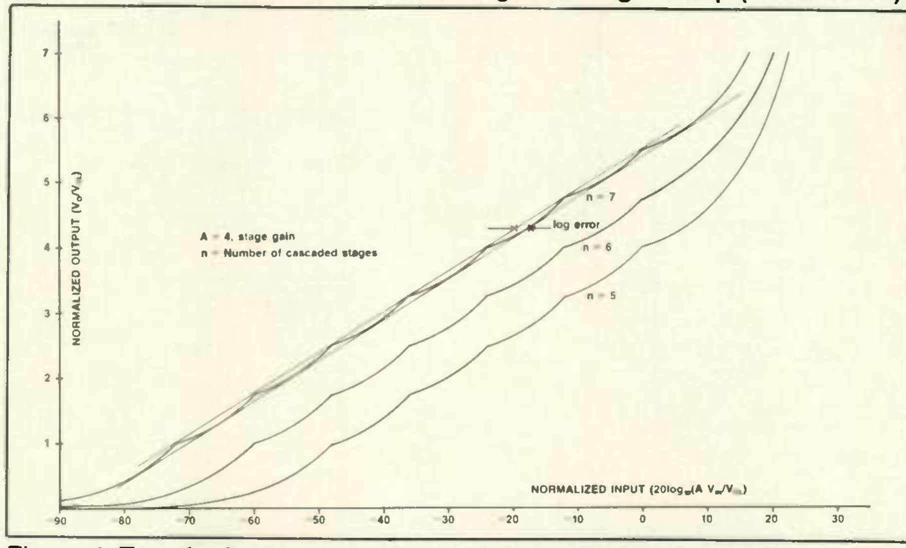

Figure 4. Transfer function of a N stage true log IF strip (Logarithmic scale).

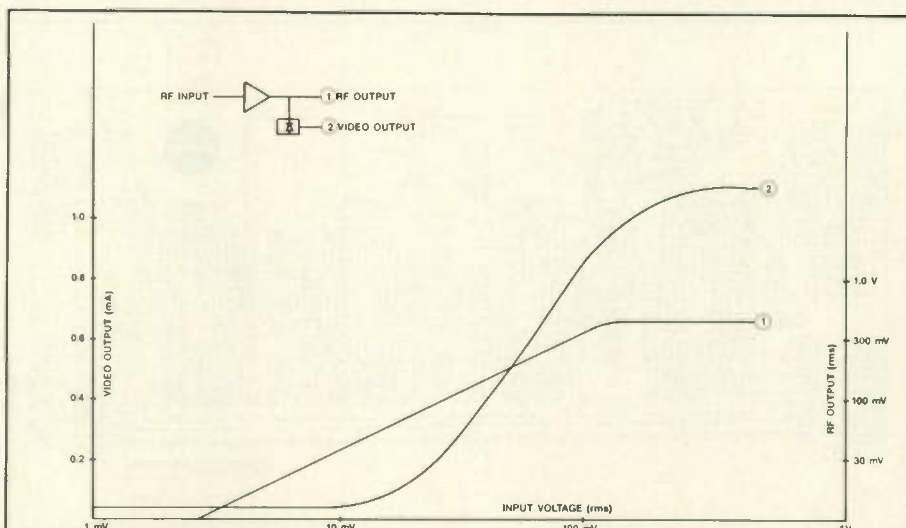

Figure 5. RF and video outputs of a log amplifier (SL521).

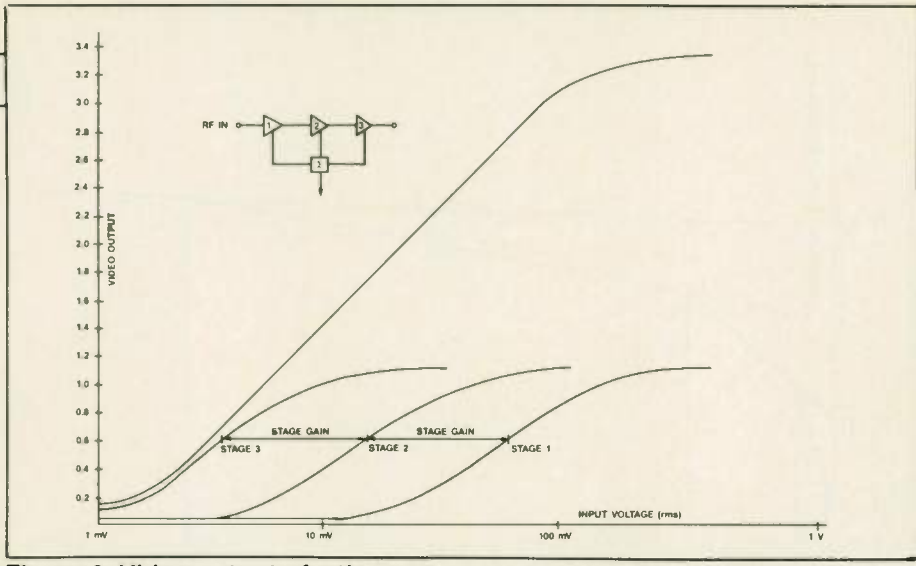

Figure 6. Video output of a three-stage successive detection log amplifier.

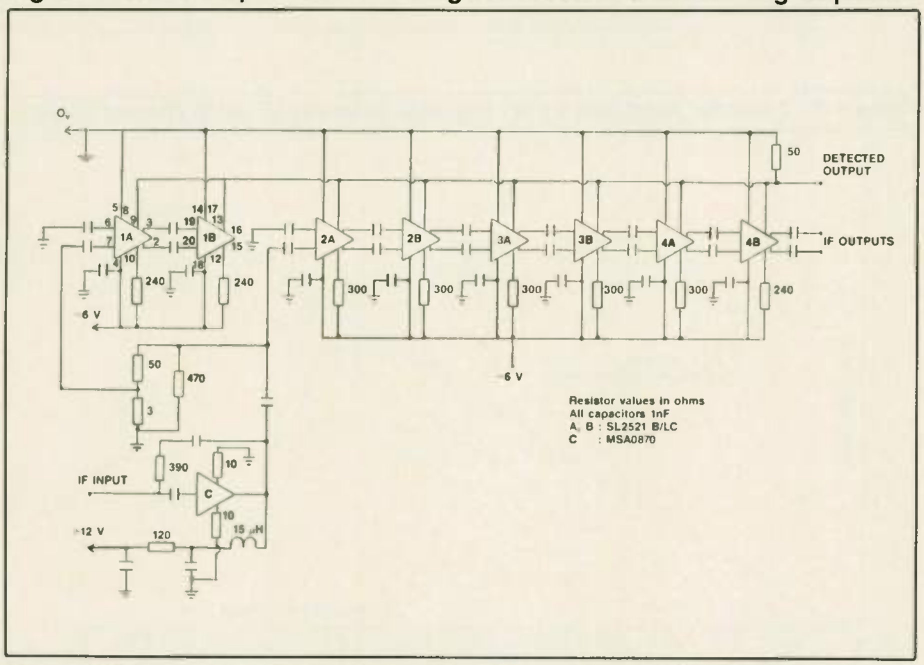

Figure 7. Circuit diagram of the eight-stage successive detection log IF strip.

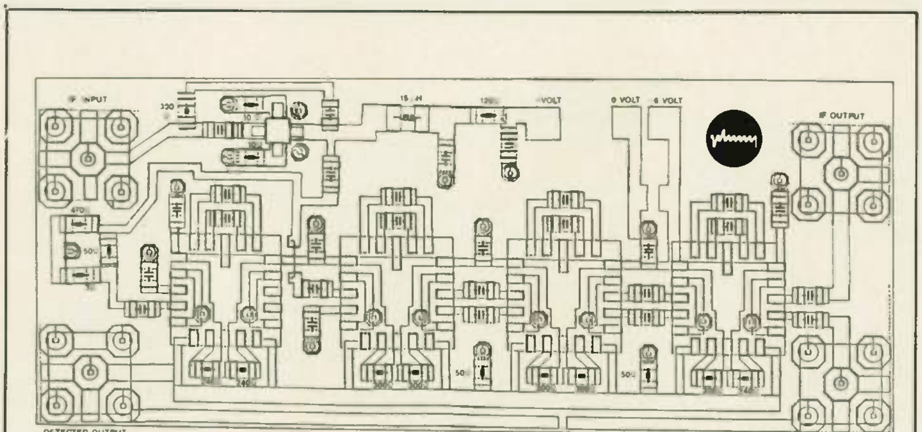

Figure 8. Component overlay of the eight-stage log IF strip.

amps are ideal for use in ultrasonic medical scanners, sonar signal amplification, and IFs for "Raycon," ECM.

Successive Detection Log Systems

Successive detection log amplifiers are the most commonly used type in military radar and ECM system applications, since they have better sensitivity than detector log video amplifiers and are more practical than true log amplifiers. A successive detection log amplifier today offers not only video output information but also IF output with phase information. For radar monopulse applications this type of log amplifier can replace individual log amplifiers and limiters which are fitted in each channel.

In order to understand the basic technique of cascading successive detection log amps, let us look at the characteristic of the SL521-log amp, shown in Figure 5. The log amp consists of a limiting amplifier with a gain A (approximately 10 dB) followed by a low level detector. Note that the video output and the RF output limit at a particular input level.

The limiting process of successive detection log amps can be accomplished by either of two ways. It can be done in the RF circuitry where the amplifiers act as limiting stages. This kind of limiting tends to be "softer" than the video limiting counterpart, with less scalloping effect on the log curve because the transition between sequential stages is smoother. The drawback is that when in limiting, those stages continue to give more output over the defined input range and may cause distortion in the desired response.

Alternatively, the limiting process can be done in the video circuitry following each detector. Since the video limiting is accomplished by a constant current limiting diode, the scalloping effect is going to be increased. This video limiting process is sometimes preferred as it provides smaller output variation over the operating temperature and better control over the log scope.

Let us consider a three-stage strip built with circuits as in Figure 5, and shown in Figure 6. The first stage of the strip gives a video output identical to the output of a single stage. The second stage receives an input signal which is increased by the gain of the first stage. This gain is constant over the range of the detector and so the second stage video output will be identical to the first, but displaced by the stage gain A. Similarly, the third stage output will be identical to the first, but displaced by two stage-gains (A).

The dynamic range of the strip can be

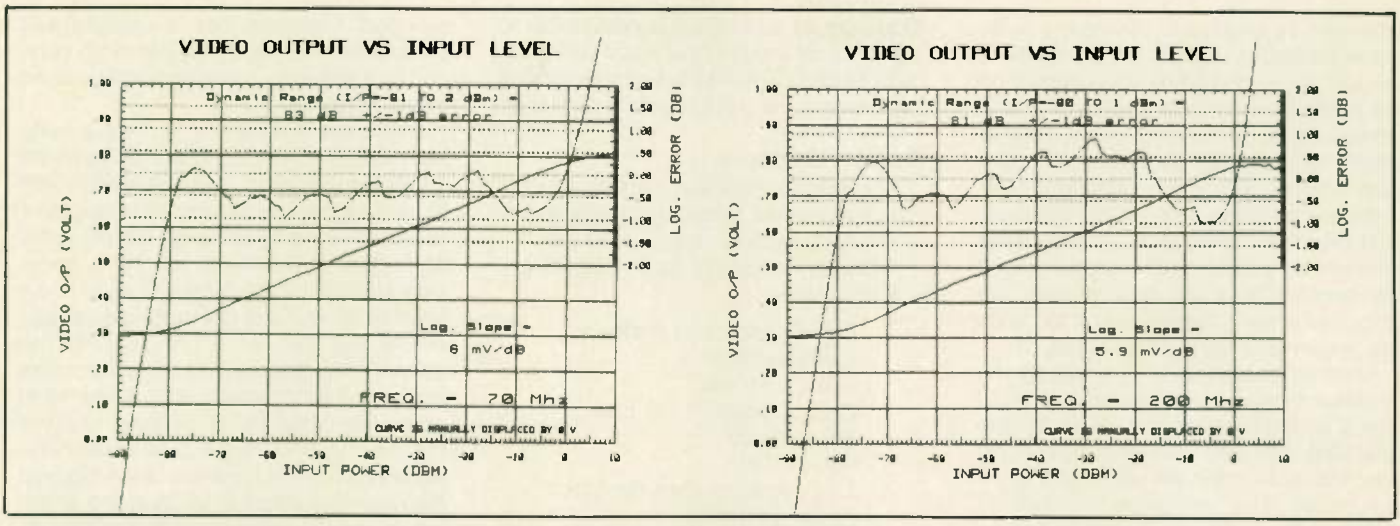

Figure 9. Detected output and logarithmic linearity at 70 and 200 MHz.

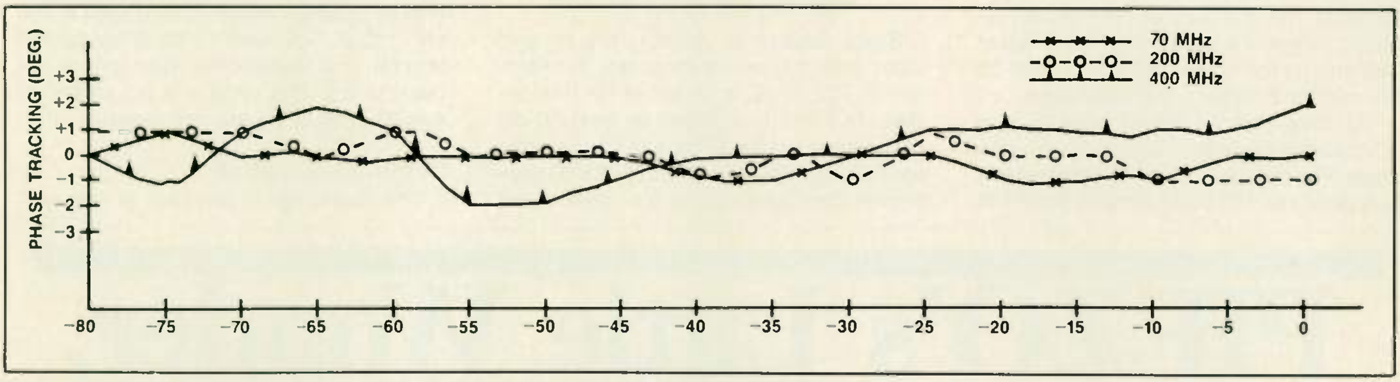

Figure 10a. Phase tracking of the two eight-stage log IF strips.

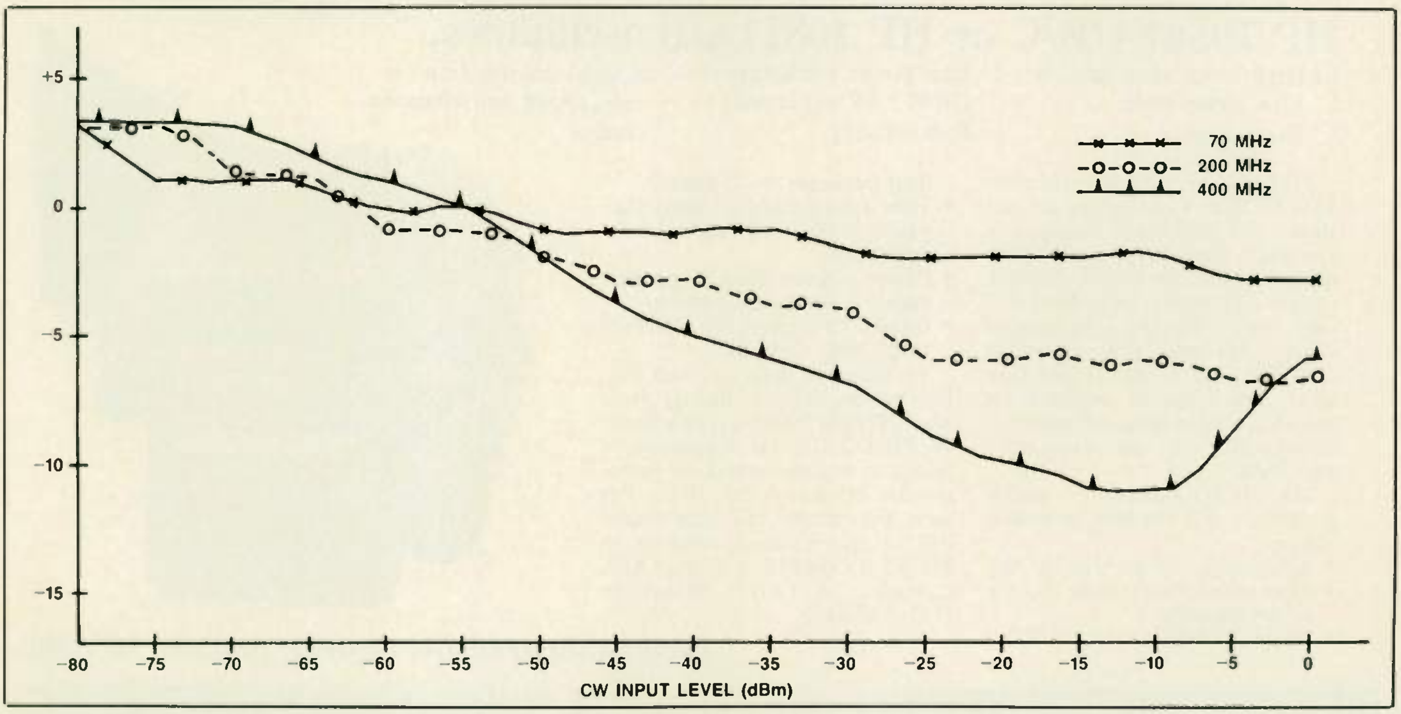

Figure 10b. Normalized phase vs. CW input level of the eight-stage log IF strip.

extended by simply increasing the number of stages, but only to the point where the last stage gives full video output on the noise produced by the first stage. If the bandwidth is reduced, the number of stages can be further increased. It is common practice to insert a bandpass filter in the center of the strip for this purpose, with the filter bandwidth a compromise between sensitivity and response time. If the bandwidth is too narrow then the response time is longer, and if too wide the system will respond to noise.

Another technique for increasing the dynamic range is to attenuate the input signal and apply it to another short strip operating in parallel with the main (longer) one. This will extend the upper end of the log curve. The maximum number of stages in the short strip is determined by the maximum input overload of the first stage of the main strip. Therefore, dynamic range of a log IF strip is limited at one end by the noise and at the other by the overload level of the first stage.

The final step in making a log IF strip is to sum the video outputs from each stage. The output of the summing network is applied to a DC coupled video amplifier. This acts as a low gain output buffer to provide a desired output slope with sufficient output current to drive the specified load resistance (typically 93Ω or lower).

Design Example

The example chosen here is a log IF strip that has the following specifications:

- Input dynamic range ⩾ 80 dB
- Center frequency up to 350 MHz
- IF output:

  Good phase and frequency characteristic.
  Gain ⩾ 65 dB.
  Output power ⩾ −5 dBm.
  NF ⩽ 10 dB.
- Video output:

  Log accuracy over dynamic range ±1.0 dB.
  20 ns pulse handling.
  Rise and fall time ⩽ 10 nsec.

Since limited IF output, phase and video information are required, the Plessey SL2521B/LC is selected for this design. In order to achieve at least 80 dB of dynamic range, 8 log amplifier stages are needed. Since each SL2521B/LC consists of 2 amplifier stages, four devices will be used. Three devices (six stages) will be cascaded in series as the main strip, and one (2 stages cascaded) will be used in the short strip.

Since the device chosen has a fairly high noise figure, a low noise preamp with 15 dB gain is used. Avantek's MSA0780 is chosen for this design. With negative feedback used to bring the 30 dB available gain down to 15 dB (with some degradation of the noise figure). Care must be taken not to put much gain ahead of the log strip or a reduction of the dynamic range could result. Input noise to the log amp, which is at some level above the threshold of the strip, reduces the net dynamic range by that amount. Added gain at the front end of the log amp also has the effect of shifting the entire input dynamic range lower. This allows extension of the dynamic range at the upper end by adding one or more stages in the short strip. However, we must be careful that the first stage of the main strip is not overloaded. The complete schematic of this IF strip is shown in Figure 7.

Circuit Description

The SL2521B/LC consists of two amplifier stages fabricated using the Plessey three-micron oxide-isolated bipolar process, packaged in a 20-pin ceramic leadless chip carrier. Each amplifier stage employs a differential amplifier with series feedback to provide linear gain. The limiting level of the amplifier is beyond the limiting level of the succeeding detector stage. The IF output signal is buffered by dual emitter followers. This output signal is also used to drive a full wave detector which determines the limiting level of the detected output. An on-chip voltage reference designed to be independent of temperature and on-chip component variation is used to provide stable operating currents for the amplfiier and detector. The SL2521B/LC also allows the detected (video) output current to be externally programmed, overcoming the problem of output current changes due to internal component variation.

The construction of the eight-stage successive detection log IF strip requires four Plessey SL2521B devices cascaded on a PC board having a ground plane on the bottom side and circuit interconnections on the top side. All components including the preamp are surface mountable except the input and output SMA connectors. Component placement is shown in Figure 8.

1 nF capacitors are used for coupling between the amplifier stages and for decoupling the power pins of each amplifier stage. 300 ohm resistors are used to program the detected output of the amplifier stages from 2A to 4A. 240 ohm resistors are used to increase the detected outputs on stages 1A, 1B and 4B and maximize the input dynamic range of the strip. The RF choke used on the bias supply of the preamp is chosen so that its impedance at the lowest operating frequency (plus the bias resistors) is at least 500 ohms. The two 10 ohm emitter feedback resistors are chosen to program the gain of the preamp. The 390 ohm feedback resistor improves the input and output VSWR of the preamp. The resistor network consisting of three resistors (470, 50 and 3 ohm) is used to match the impedance of the preamp and to provide the proper amount of attenuated signal to the short strip. The detected outputs of the smaller stages are tied together with a 50 ohm pull up resistor and are brought out to the SMA connector.

Typical Performance

The eight-stage log IF strip which consists of four SL2521 log amps in cascade and one MSA0870 preamp has been evaluated, with the following results:

Output power, gain and *bandwidth* were measured at one IF output port with the other port terminated into 50 ohms. The limited output power and small signal gain were −4.7 dBm and 65 dB, respectively at 200 MHz; and −4.9 dBm and 64 dB, respectively at 400 MHz. 3 dB bandwidth was 480 MHz.

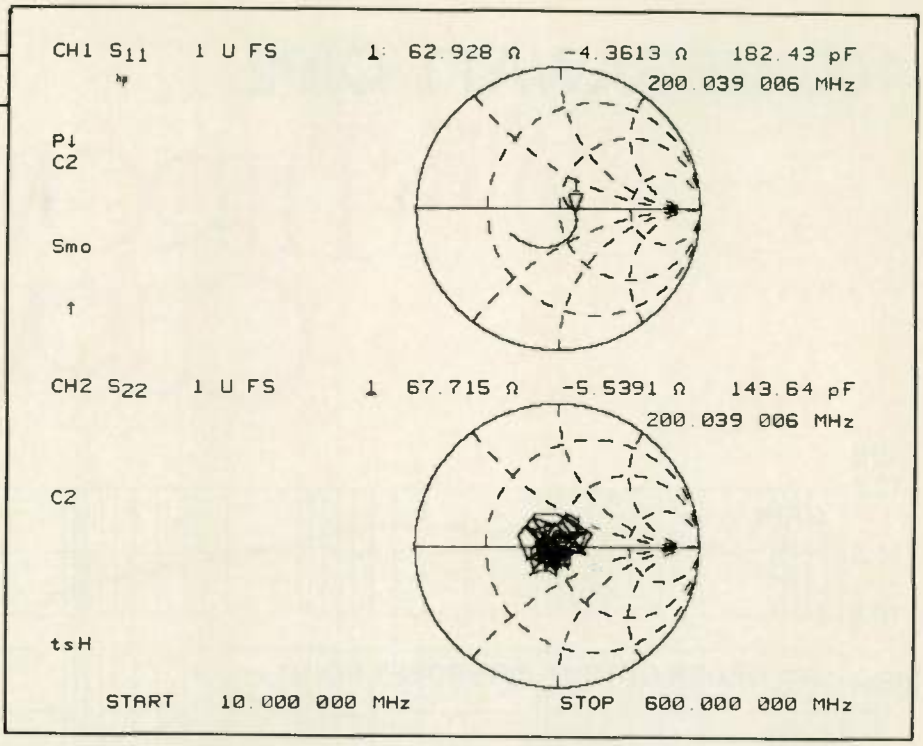

Figure 11. IF input and output reflection coefficients of the eight-stage log IF strip.

Dynamic range at the video output port was measured under CW conditions, and as shown in Figure 9, 83 dB of dynamic range was obtained at 70 MHz, and 81 dB at 200 MHz with ±1 dB linearity.

Phase shift at the IF output under CW conditions was less than 6° at 70 MHz, 10° at 200 MHz, and 14° at 400 MHz (Figure 10b). For the phase tracking measurement, a second log strip was built with devices from different lots. As shown in Figure 10a, the two strips track within ±1° at 70 MHz and 200 MHz, and within ±2° at 400 MHz over the full 80 dB input range.

Noise figure was 7.5 dB at 70 MHz, 8.0 dB at 200 MHz, and 8.5 dB at 400 MHz.

VSWR, S11 and *S22* reflection coefficients (Figure 11) were measured at an input level of −30 dBm, with an input VSWR of 1.2:1 at 70 MHz, 1.3:1 at 200 MHz, and 1.6:1 at 400 MHz. Output VSWR was 1.3:1 at 70 MHz, 1.5:1 at 200 MHz, and 1.2:1 at 400 MHz.

The *IF limiting* characteristics were examined in some detail. Good limiting was achieved at various frequencies, with good symmetry achieved with the aid of filtering.

The *pulse output* of the video output port was examined and a 20 ns pulse at 0 dBm of input signal level demonstrated good rise and fall times (less than 5 ns).

Among the three types of log amps, successive detection log amps are the most commonly used in airborne radar applications. They provide not only video information but also phase information and phase tracking capability. These characteristics can be achieved with monolithic log amps where special bipolar process and small geometry are employed. In their hybrid counterparts, where discrete transistors and diodes are used, both video and phase information cannot be achieved simultaneously.

It is also worth noting that monolithic log amps have most performance characteristics inherent in their design, whereas hybrid log amps have to be individually optimized and adjusted for each characteristic, often resulting in higher costs and larger size. rf

References

(1) D. Potson and R.S. Hughes, "Extremely Fast Log Amp Handles Narrow Pulses." *Microwaves and RF*, Part 1, April 1985, pp.85-90, 150.

(2) W.L. Barber and E.R. Brown, "A True Logarithmic Amplifier for Radar IF Applications," *IEEE Journal of Solid State Circuit Design*, June 1980.

About the Author

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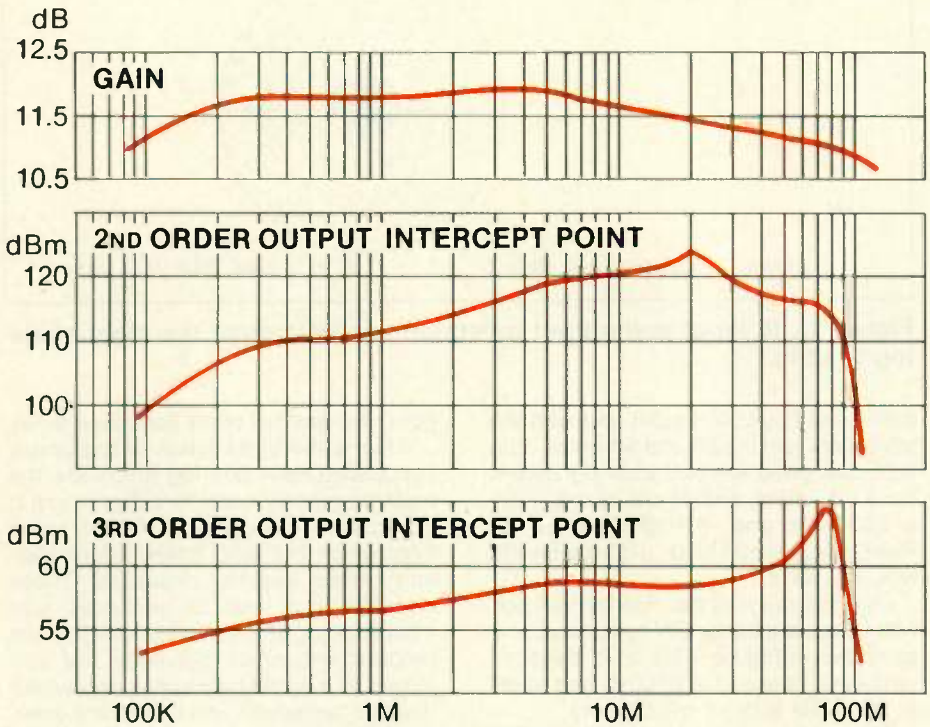

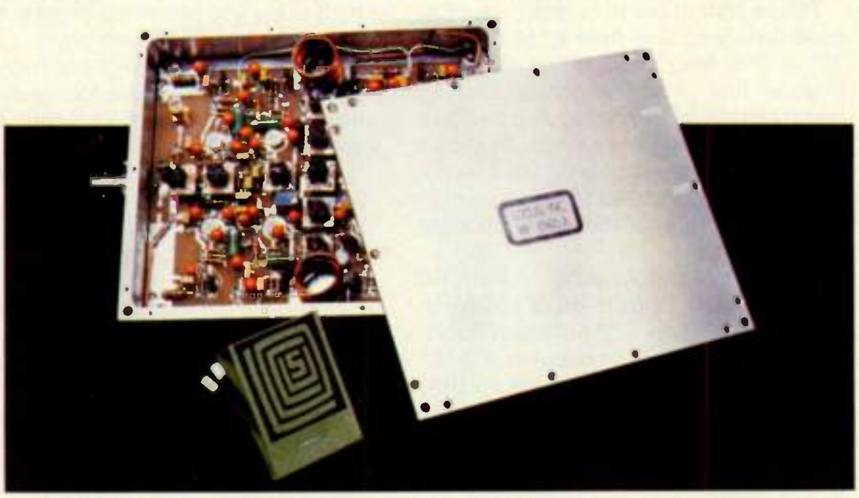

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Constant Impedance Bandpass and Diplexer Filters

By Francois Methot
Le Groupe Videoway Inc.

Constant impedance bandpass filters and diplexer filters are all used to eliminate most problems related to circuit design of mixers, limiters and phase comparators: poor isolation (LO-IF), unwanted beats, saturation of amplifying stages at mixer output, DC offset, etc. This article illustrates a technique used to calculate certain types of constant impedance filters (Butterworth, Chebyshev) using normalized value tables in conjunction with a calculation program developed for the IBM PC and the HP-41C calculator.

In a mixer, conversion of an IF frequency into a baseband produces at the mixer output unwanted frequencies such as F(LO)-F(IF), F(LO), F(LO)+F(IF) and others. In a system with an IF lower than the input RF and LO, a low-pass filter easily eliminates most unwanted frequencies, but will present an impedance which is too high or too low outside the operating band. Due to this poor matching, spurious outputs return inside the mixer and cause additional beats (1).

These beats can be eliminated by ensuring that all mixer inputs and outputs are sourced or terminated with the same impedance as the mixer (typically 50Ω or 75Ω). A filter must then be calculated that will feed the mixer a constant impedance for the entire frequency range of operation. Bandpass and diplexer filters will be described that include this feature (Figures 1 and 2).

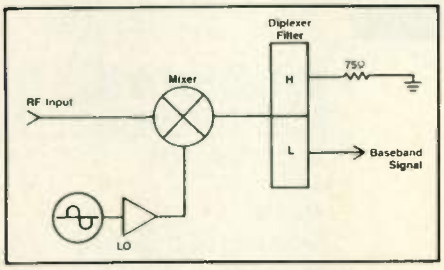

Figure 1. Basic mixer diplexer configuration.

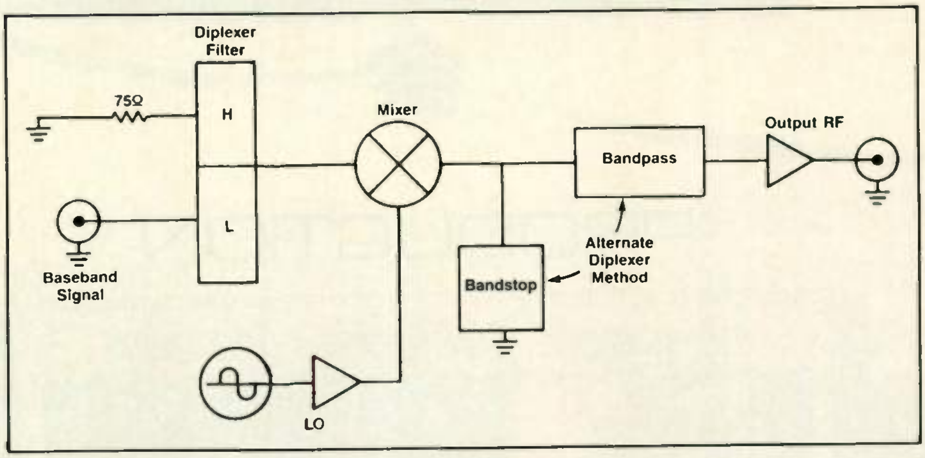

Figure 2. Impedance matching with diplexer.

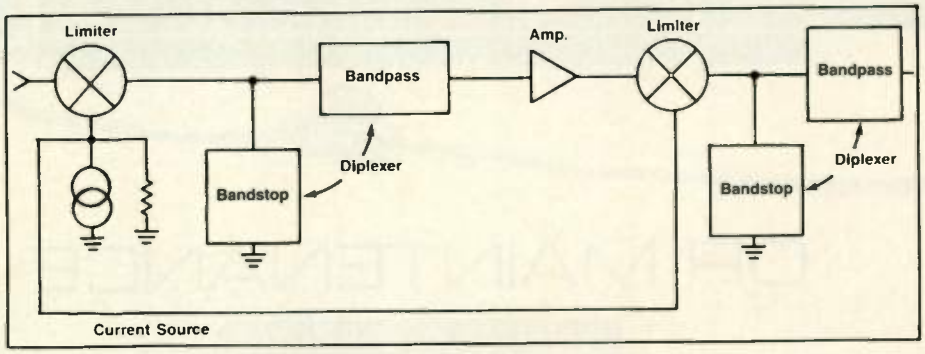

Figure 3. Diplexers in a limiter application.

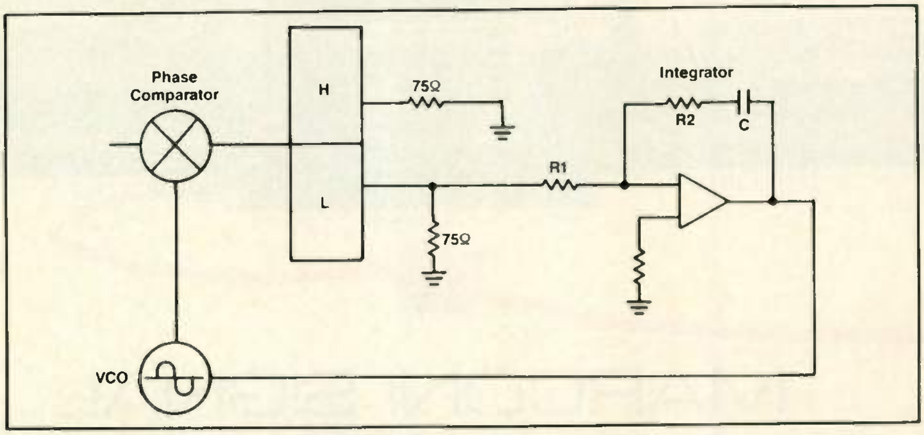

Figure 4. PLL phase comparator application.

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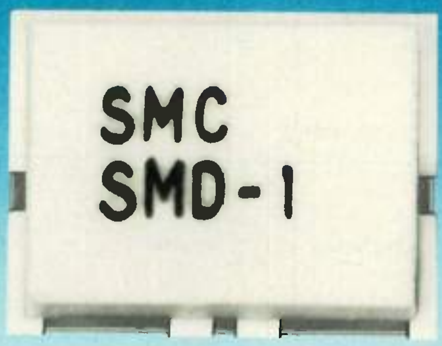

.375 x .500 x .150

MODEL	RF/LO MHz	IF MHz	LO POWER dBm
SMD-1	1-1000	DC-1000	+7
SMD-1M	1-1000	DC-1000	+17
SMD-1H	1-1000	DC-1000	+23
SMD-2	20-1500	DC-1500	+7
SMD-2M	20-1500	DC-1500	+17
SMD-2H	20-1500	DC-1500	+23
SMD-3	20-2500	20-600	+7
SMD-3M	20-2500	20-600	+17
SMD-3H	20-2500	20-600	+23

.375 x .500 x .140

MODEL	RF/LO MHz	IF MHz	LO POWER dBm	PRICE (100 pcs.)
SMD-C1	1-1000	DC-1000	+7	$ 6.50
SMD-C1M	1-1000	DC-1000	+17	7.00
SMD-C1H	1-1000	DC-1000	+23	11.00
SMD-C2	20-1500	DC-1500	+7	7.00
SMD-C2M	20-1500	DC-1500	+17	8.00
SMD-C2H	20-1500	DC-1500	+23	12.00
SMD-C3	20-2500	20-600	+7	8.90
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Other Circuits and Applications

The technique outlined above can also be applied to reduce additional beats generated by mixers used as limiters (Figure 3) or as phase comparators (Figure 4). In a phase locked loop (PLL), harmonic distortions produced by mixers used as saturation phase comparators should be filtered before amplification of the DC correction of the PLL. To avoid dedgradation of the PLL operation, it is important to select a cut off point 10 times higher than the phase locked loop's natural frequency ωN).

As shown in Figure 5, when a frequency must be substituted for another, bandpass filters and band stop filters can be used to trap out the unwanted signal and insert a new one. This technique is common in the CATV industry.

Another application is in circuits requiring bidirectional amplifiers. One must arrange diplexer filters in an order that will give an isolation higher than the power gain of amplifiers in order to avoid oscillation problems (Figure 6).

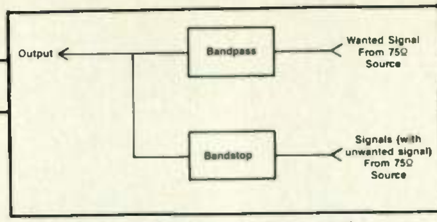

Figure 5. "Trap and insert" application.

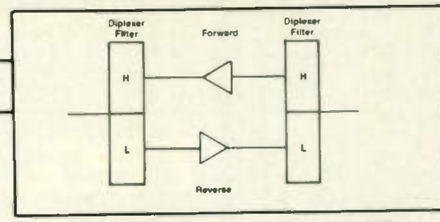

Figure 6. Bidirectional amplifier system.

Filter Calculation Technique

In theory, it is possible to design perfect diplexer filters by selecting normalized values (singly terminated) among elements from Butterworth and maximally flat filter tables (2). For instance, if a low-pass filter is combined with a high-pass filter calculated according to those tables, the sum of the impedance and the admittance combines as illustrated in Figure 7.

The voltage source model is as follows:

$Y_{in} = Y_{in}$ (low pass) + Y_{in} (high pass) = 1

Therefore,

$R_e[Y_{in}(\text{low pass})] + R_e[Y_{in}(\text{high pass})] = 0$

and,

$I_m[y_{in}(\text{low pass})] + I_m[Y_{in}(\text{high pass})] = 0$

The same results can be obtained with the current source model.

However, if Chebyshev filters are used, the elements' normalized values are no longer accurate because the sum of impedances and admittances is no longer equal to 1. In that case, the normalized values of Chebyshev filters must be modified by 1.023 factor to produce, at the cut off point, a value of 0.5 impedance or admittance for each high-pass or low-pass filter(3).

Since Chebyshev filters have a better selectivity and it is possible to improve their input return loss, we have included them in the calculation program at the end of this article. Moreover, the reader may refer to several normalized and modified value tables suitable for other Chebyshev filters (0.25 dB to 1 dB) as mentioned by Veltrop and Wilds(3).

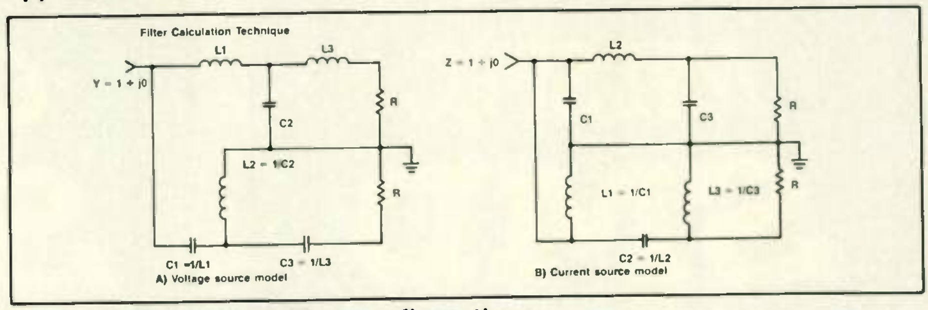

Figure 7. Normalized filter configurations.

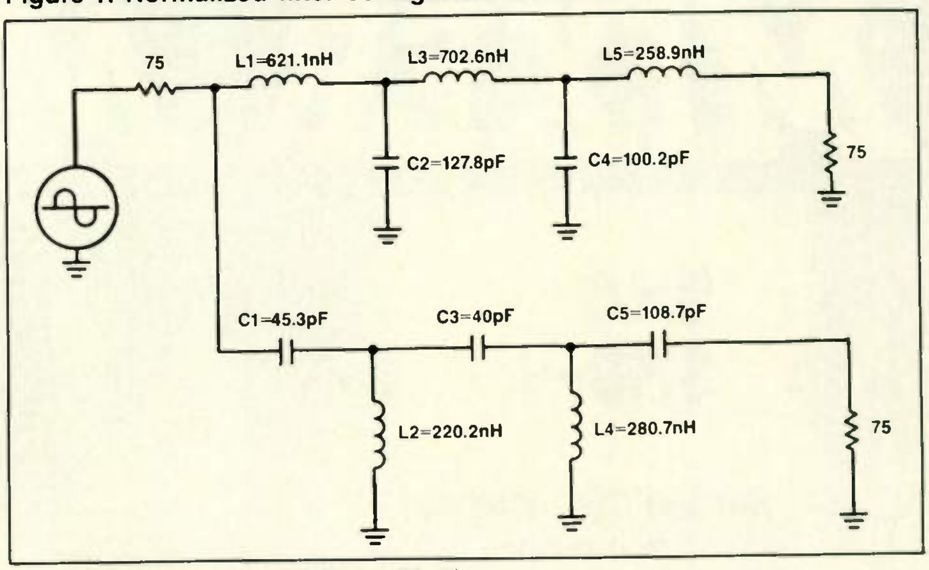

Figure 8(a). Example diplexer (N=5).

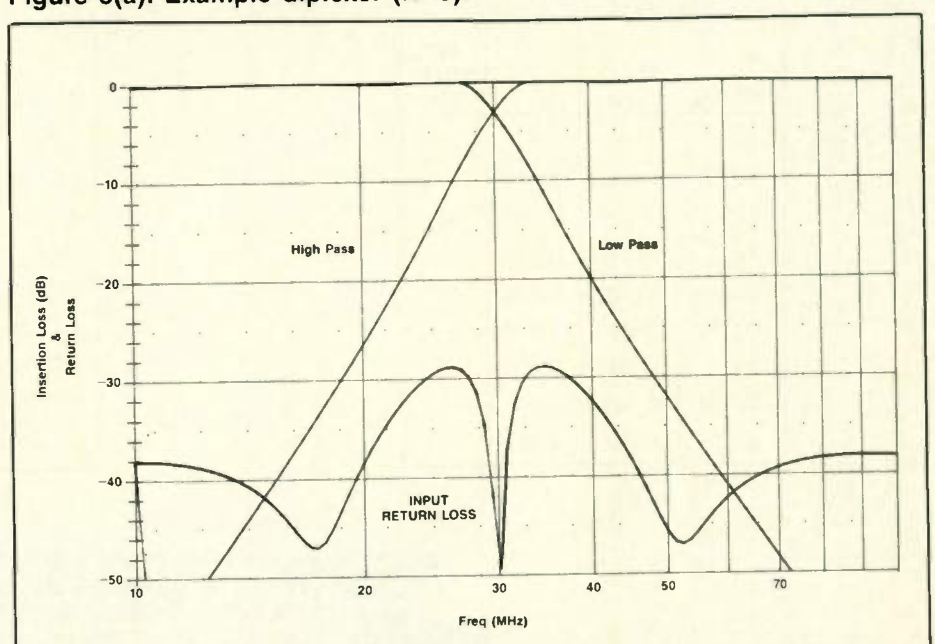

Figure 8(b). Performance curves for highpass/lowpass diplexer.

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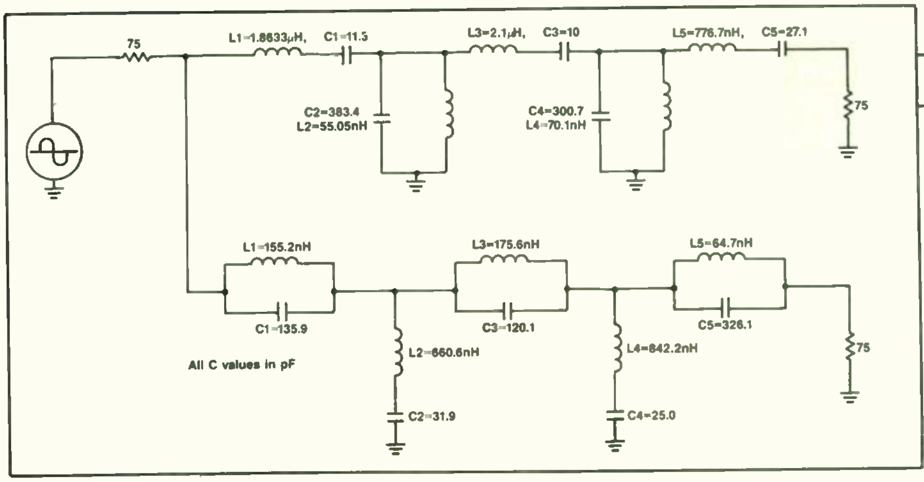

Figure 9(a). Example bandpass/bandstop diplexer.

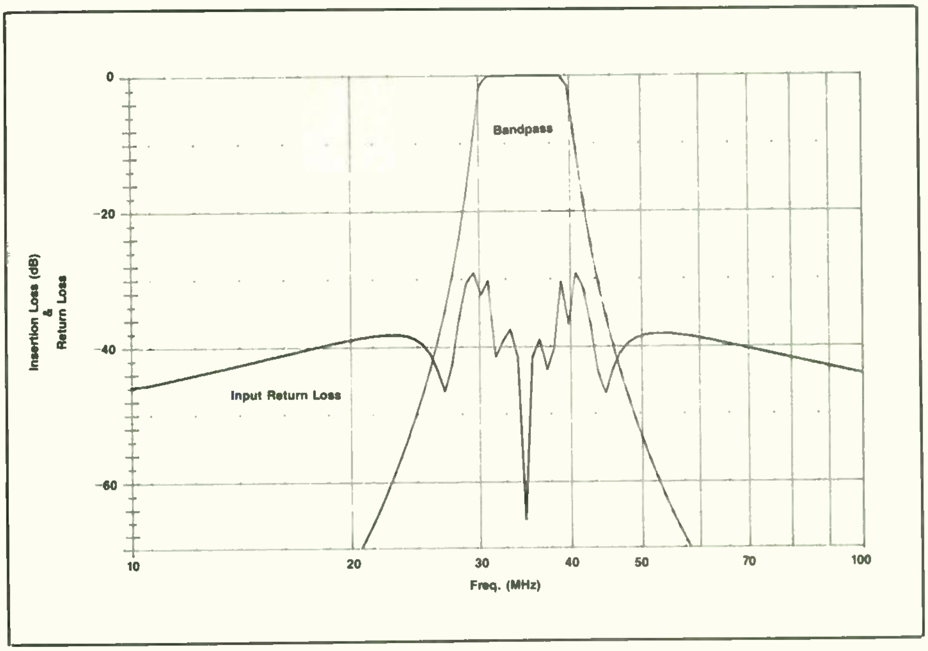

Figure 9(b). Performance curves for bandpass/bandstop diplexer.

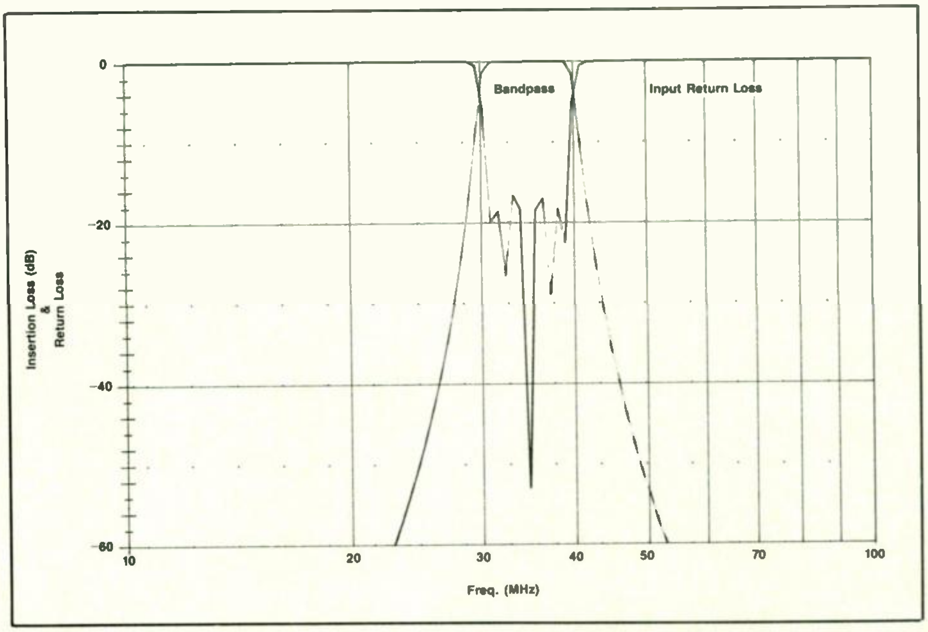

Figure 10. Demonstration of poor matching outside filter passband.

```
run
(1) For Low/High Pass, (2) For Band Pass/Stop? 1
Order 3, 5, or 7? 5
Impedance Z? 75
Cut Off Frequency? 30e6
       ** Low Pass Section **
L 1=  621.10n      C 2=  127.81p
L 3=  702.62n      C 4=  100.25p
L 5=  258.90n
          ** High Pass Section **
C 1=  45.314p      L 2=  220.20n
C 3=  40.056p      L 4=  280.73n
C 5=  108.70p
New calculation (Y/N)? Y

(1) For Low/High Pass, (2) For Band Pass/Stop? 2
Order 3, 5, or 7? 5
Impedance Z? 75
          ** Band Pass / Band Stop **
Low Frequency? 30E6
High Frequency? 40E6
             ** Band Pass Section **
C 1=  11.328p    L 1=  1.8633u     Serial Con.  Serial Res.
C 2=  383.43p    L 2=  55.051n     Parall Con.  Parall Res.
C 3=  10.014p    L 3=  2.1078u     Serial Con.  Serial Res.
C 4=  300.76p    L 4=  70.183n     Parall Con.  Parall Res.
C 5=  27.176p    L 5=  776.71n     Serial Con.  Serial Res.
             ** Band Stop Section **
C 1=  135.94p    L 1=  155.27n     Serial Con.  Parall Res.
C 2=  31.952p    L 2=  660.61n     Parall Con.  Serial Res.
C 3=  120.16p    L 3=  175.65n     Serial Con.  Parall Res.
C 4=  25.063p    L 4=  842.20n     Parall Con.  Serial Res.
C 5=  326.12p    L 5=  64.726n     Serial Con.  Parall Res.
New calculation (Y/N)? N

Ok
```

Figure 11. Sample computations using diplexer design program.

```
10 'PROGRAM DIPLEXER
20 ' Jean-Luc Dugas MAY 15 86
30 DIM TABLE(7)
40 INPUT "(1) For Low/High Pass, (2) For Band Pass/Stop";A
50 IF A<>1 AND A<>2 THEN 40
60 INPUT "Order 3, 5, or 7";ORDER
70 DATA 1.5133,1.509,.7164
80 DATA 1.561,1.8069,1.7659,1.4173,.6507
90 DATA 1.5748,1.8577,1.921,1.827,1.734,1.3786,.6307
100 IF ORDER=3 THEN RESTORE 70 ELSE IF ORDER = 5 THEN RESTORE 80 ELSE IF ORDER =
 7 THEN RESTORE 90 ELSE 60
110 INPUT "Impedance Z";Z
120 FOR N=1 TO ORDER
130 READ TABLE(N)          'Read normalized values
140 NEXT N
150 ON A GOSUB 210,480     'Select Filter type
160 INPUT "New calculation (Y/N)";YES$
170 PRINT
180 IF YES$<>"N" THEN 40
190 END
200 '** Low Pass  High Pass Section **
210 INPUT "Cut Off Frequency";FC
220 WC=6.28319*FC
230 PRINT TAB(10) "** Low Pass Section **
240 FOR N=1 TO ORDER STEP 2
250 TEST = TABLE(N)*Z/WC          'L(N)
260 GOSUB 940
270 LIN$="L"+STR$(N)+TEST$
280 IF N=ORDER THEN 320
290 TEST=TABLE(N+1)/(Z*WC)         'C(N+1)
300 GOSUB 940
310 LIN$=LIN$+"     C"+STR$(N+1)+TEST$
320 PRINT LIN$                    'Output results of L(N), C(N+1)
330 NEXT N
340 PRINT TAB(10),"** High Pass Section **"
350 FOR N=1 TO ORDER STEP 2
360 TEST=1/(TABLE(N)*WC*Z)         'C(N)
370 GOSUB 940
380 LIN$="C"+STR$(N)+TEST$
390 IF N= ORDER THEN 430
400 TEST=Z/(WC*TABLE(N+1))          'L(N+1)
410 GOSUB 940
420 LIN$=LIN$+"     L"+STR$(N+1)+TEST$
430 PRINT LIN$                       'Output results C(N), L(N+1)
440 NEXT N
450 RETURN
460 '
470 '
480 PRINT TAB(10)"** Band Pass / Band Stop **"
490 INPUT "Low Frequency";F1
500 W1=6.28319*F1
510 INPUT "High Frequency";F2
520 W2=6.28319*F2
530 PRINT TAB(10),"** Band Pass Section **"
540 FOR N= 1 TO ORDER STEP 2
550 TEST=(F2-F1)/(W1*F2*Z*TABLE(N))                        'C(N)
560 GOSUB 940
570 LIN$="C"+STR$(N)+TEST$
580 TEST=Z*TABLE(N)/(W2-W1)                               'L(N)
590 GOSUB 940
600 LIN$=LIN$+"   L"+STR$(N) + TEST$ + "     Serial Con.  Serial Res."
610 PRINT LIN$                                             'Output Results C(N), L(N)
620 IF N= ORDER THEN 700
630 TEST=TABLE(N+1)/((W2-W1)*Z)                           'C(N+1)
640 GOSUB 940
650 LIN$="C" + STR$(N+1) + TEST$
660 TEST=Z*(F2-F1)/(W1*F2*TABLE(N+1))                     'L(N+1)
670 GOSUB 940
680 LIN$= LIN$+"   L"+STR$(N+1)+TEST$+"     Parall Con.  Parall Res."
690 PRINT LIN$                                             'Output Results C(N+1), L(N+1)
700 NEXT N
710 PRINT TAB(10),"** Band Stop Section **"
720 FOR N= 1 TO ORDER STEP 2
730 TEST = 1/(TABLE(N)*(W2-W1)*Z)                         'C(N)
740 GOSUB 940
750 LIN$="C"+STR$(N)+TEST$
760 TEST=(F2-F1)*TABLE(N)*Z/(W2*F1)                       'L(N)
770 GOSUB 940
780 LIN$=LIN$+"   L"+STR$(N)+TEST$+"     Serial Con.  Parall Res."
790 PRINT LIN$                                             'Output Results C(N), L(N)
800 IF N= ORDER THEN 880
810 TEST=TABLE(N+1)*(F2-F1)/(W2*F1*Z)                     'C(N+1)
820 GOSUB 940
830 LIN$="C"+STR$(N+1)+TEST$
840 TEST=Z/(TABLE(N+1)*(W2-W1))                           'L(N+1)
850 GOSUB 940
860 LIN$=LIN$+"   L"+STR$(N+1)+TEST$+"     Parall Con.  Serial Res."
870 PRINT LIN$                                             'Output Results C(N+1), L(N+1)
880 NEXT N
890 RETURN
900 '
910 '
920 'Eng notation and string manipulation
930 '
940 I=0
950 WHILE TEST<=1 AND I<5
960 TEST=TEST*1000
970 I=I+1
980 WEND
990 IF I=0 THEN 1030
1000 DENOM$="munpf"
1010 TEST$="= "+LEFT$(STR$(TEST),7)+MID$(DENOM$,I,1)+"  "
1020 RETURN
1030 TEST$="= "+LEFT$(STR$(TEST),7)+"?  "        'Failure to find Eng Not.
1040 RETURN
```

BASIC program for computation of diplexer filter component values.

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T-2S63U-60	2-4	20	0.5	1.25:1
T-4S63U-50	4-8	20	0.4	1.25:1
T-7S43U-30	7-11	28	0.4	1.10:1
T-7S83U-40	7.6-18	16	0.8	1.50:1
T-8S43U-20	8-12.4	18	0.4	1.30:1
T-8S63U-30	8-16	17	0.5	1.35:1
T-10S63U-20	10-20	17	0.7	1.35:1
T-12S43U-30	12-18	18	0.5	1.30:1
T-18S33U-20	18-26.5	17	1.0	1.50:1

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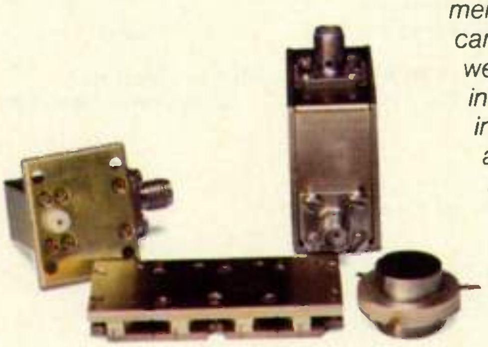

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INFO/CARD 73

Using Normalized Tables of Filters

The value of the bandpass and diplexer filter components can be calculated using normalized tables for Butterworth and Chebyshev low-pass filters according to the following procedure:

1) Select in the row of *RS=infinite*, those normalized values corresponding to the voltage source configuration.

$$\text{RS becomes} \quad \frac{1}{RS} = 0\Omega$$

2) Invert low-pass component value, thus inductances become capacitances and capacitances inductances (Figures 7(a) and 7(b)). The high-pass filter normalized component values are obtained through this conversion.

3) In the case of diplexer filters, normalized values must then be modified according to selected frequency and impedance by using the following formulas:

$$L = \frac{L'R}{\omega}$$

$$C = \frac{C'}{\omega R}$$

4) To make a constant impedance bandpass filter, one must simply transform the low-pass model into a bandpass and a band stop(4). Figures 9a and 9b demonstrate this configuration. There are the drawbacks that this type of filter requires a greater number of components (double), and the filters must be accurately fine tuned to optimize the input return loss.

Using the Calculation Programs

It is also possible to calculate the value of the components using the calculation programs developed for the IBM PC and the HP-41C calculator described in this article. Both programs compute the components' modified values for Chebyshev type filters with .1 dB of ripple and for the orders N=3, 5 and 7.

Two constant impedance filter sample computations are included in Figure 11. These are the designs shown in Figures 8 and 9. Figure 10 illustrates the curves of a simple bandpass filter to demonstrate that the return loss outside the band is not adequate to use it with mixers.

Acknowledgement

The author wishes to thank Jean-Luc Dugas for developing the HP-41 calculator and BASIC calculation programs.

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About the Author

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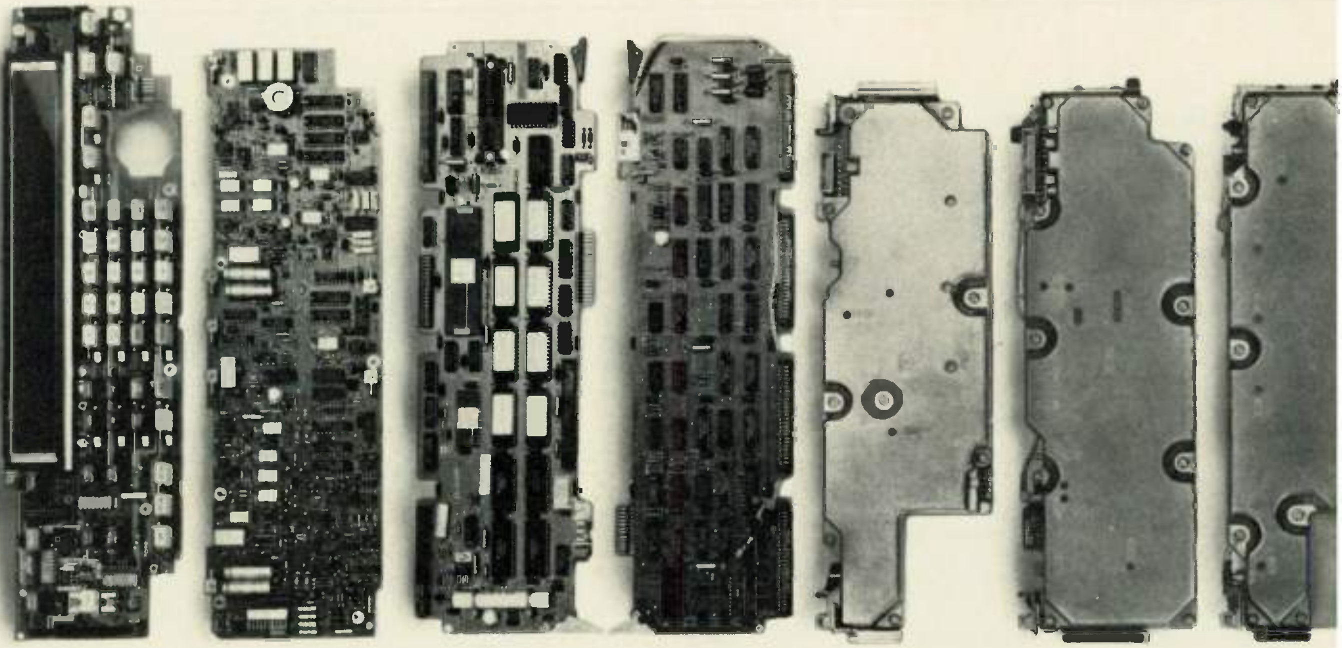
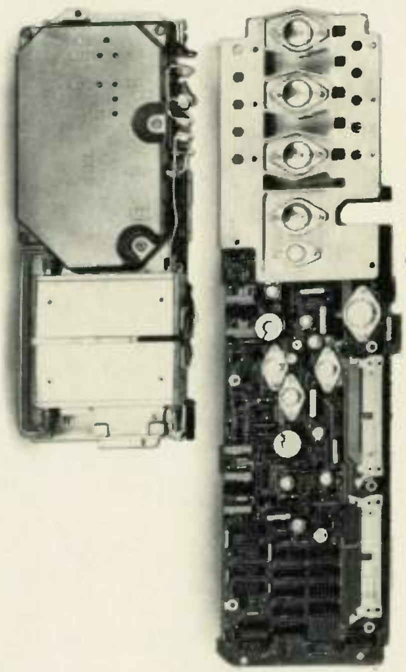
```
LBL'DIPLEX
CF 01
CF 02
'S=
ASTO 29
'P=
ASTO 30
29.03001
STO 28
'CHOOSE ORDER
AVIEW
PSE
'CHEBY.1dB
AVIEW
PSE
'N=3, 5 OR 7
PROMPT
3
X <> Y
X = Y ?
XEQ 'N=3
5
X <> Y
X = Y ?
XEQ 'N=5
7
X = Y ?
XEQ 'N=7
'Z ?
PROMPT
STO 10
'BP/BS=0 LP/HP=
PROMPT
X = 0 ?
GTO 03
'LOW PASS
AVIEW
PSE
'FREQ ?
PROMPT
2
*
PI
*
STO 11
RCL 10
*
1/X
STO 12
RCL 10
RCL 11
/
STO 13
LBL 00
'L
SF 02
XEQ 10
RCL IND 00
RCL 13
*
XEQ 11
ISG 00
GTO 01
GTO 02
LBL 01
'C
SF 02
XEQ 10
RCL IND 00
RCL 12
*
XEQ 11
ISG 00
GTO 00
LBL 02
FS ? 01
GTO DIPLEX
'HIGH PASS
AVIEW
XEQ 12
SF 01
RCL 28
1
-
STO 28
GTO 01
LBL 03
'F1 <LOW> ?
PROMPT
STO 20
'F2<HI> ?
PROMPT
STO 21
X <> Y
-
STO 22
RCL 21
RCL 20
*
STO 23
2
*
PI
*
STO 24
RCL 10
*
STO 26
2
PI
*
RCL 22
*
STO 25
RCL 10
*
STO 27
LBL 04
'BAND PASS
AVIEW
PSE
XEQ 05
'BAND STOP
AVIEW
XEQ 12
XEQ 06
GTO 'DIPLE
LBL 05
'C
XEQ 10
RCL 22
RCL 26
RCL IND 00
*
/
SF 02
XEQ 11
'L
XEQ 10
RCL 10
RCL IND 00
*
RCL 25
/
XEQ 11
ISG 00
GTO 06
RTN
LBL 06
'C
XEQ 10
RCL IND 00
RCL 27
/
SF 02
XEQ 11
'L
XEQ 10
RCL 10
RCL 22
*
RCL 24
RCL IND 00
*
/
XEQ 11
ISG 00
GTO 05
RTN
LBL 10
FIX 0
ARCL 00
ARCL IND 2
FS?C 28
ISG 28
RTN
29.03001
STO 28
RTN
LBL 11
ENG 3
ARCL X
PROMPT
RTN
LBL 12
RCL 00
FRC
STO 00
ISG 00
LBL 13
RCL IND 00
1/X
STO IND 00
ISG 00
GTO 13
RCL 00
FRC
STO 00
ISG 00
RTN
LBL 'N=3
1.00301
STO 00
1.5128
STO 01
1.5085
STO 02
.7162
STO 03
RTN
LBL 'N=5
1.00501
STO 00
1.5614
STO 01
1.8072
STO 02
1.7661
STO 03
1.4174
STO 04
.6508
STO 05
RTN
LBL 'N=7
1.00701
STO 00
1.5748
STO 01
1.8578
STO 02
1.921
STO 03
1.827
STO 04
1.734
STO 05
1.3786
STO 06
.6308
STO 07
RTN
.END.
```

HP-41 version of diplexer filter design program.





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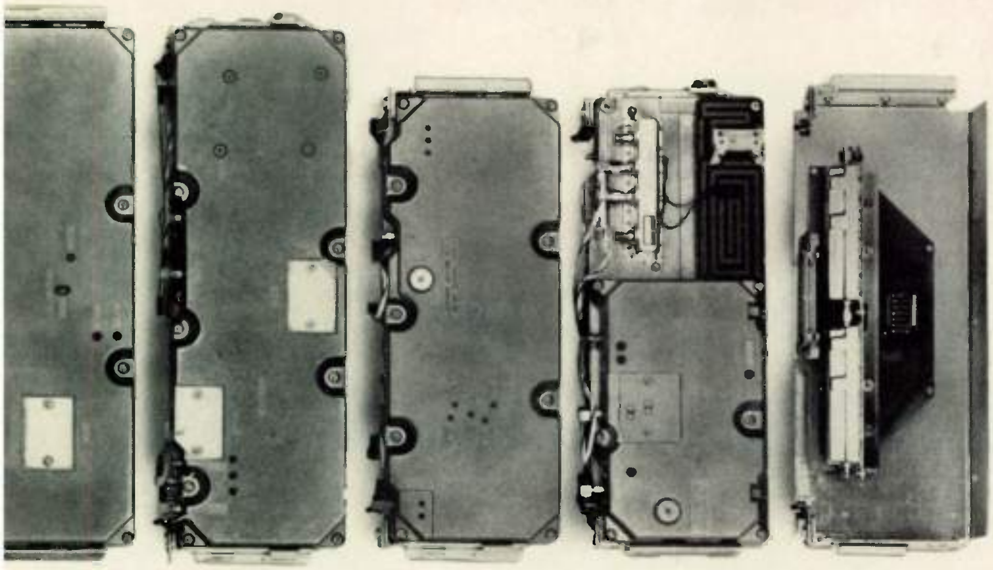

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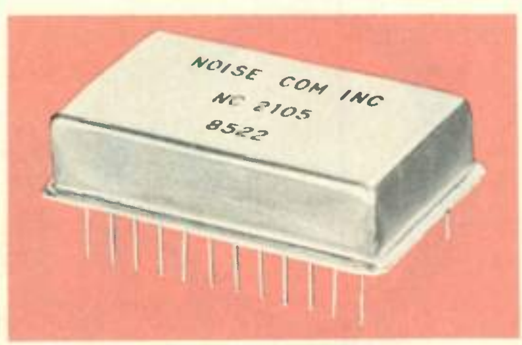

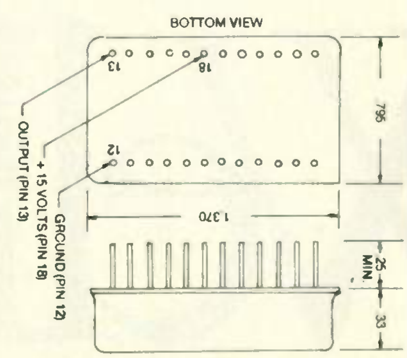

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Broadband Noise Improvement in RF Power Amplifiers

Reducing RFI in Multiple Transmitter/Receiver Applications.

By Ernie Franke and Joseph DeLeon
E-Systems ECI Division

Operation of a high power transmitter in the vicinity of a sensitive receiver can result in the degradation of a receiver's performance due to broadband transmitter noise. This includes both duplex operation (simultaneous transmission and reception on two different channels using a single antenna) and co-location operation (multiple receivers operating in the vicinity of a high power transmitter). This broadband noise may originate in a synthesizer or exciter or may even be dominated by the noise figure of the power amplifier chain itself. The noise spectrum may stretch for several megahertz on both sides of the carrier. The broadband noise of present wideband transistor amplifiers is greater than was present in older tube-type amplifiers which contained tuned circuits or cavity output networks. The low-Q, untuned interstage and output networks present in solid-state transmitters increase the available bandwidth and thus increase the output noise power on the receiver frequency.

Broadband transmitter noise adversely affects the performance of a receiver. Ideally, the transmitter should confine all of its output power within a narrow band of frequencies on either side of the transmit frequency. The bulk of the power is in fact confined within the assigned transmitter channel but undesired broadband noise exists on frequencies above and below the carrier frequency. Filter circuits in the transmitter eliminate a considerable portion of the undesired radiation but enough noise energy is conducted to the antenna to degrade the performance of a receiver operating several MHz away. The level of noise is greatest at frequencies close to the carrier frequency of the transmitter. Transmitter noise appears as "on-channel" noise interference to the receiver and cannot be filtered out at the receiver. It falls exactly on the operating frequency of the receiver and competes with the desired signal. Assuming that the frequency separation

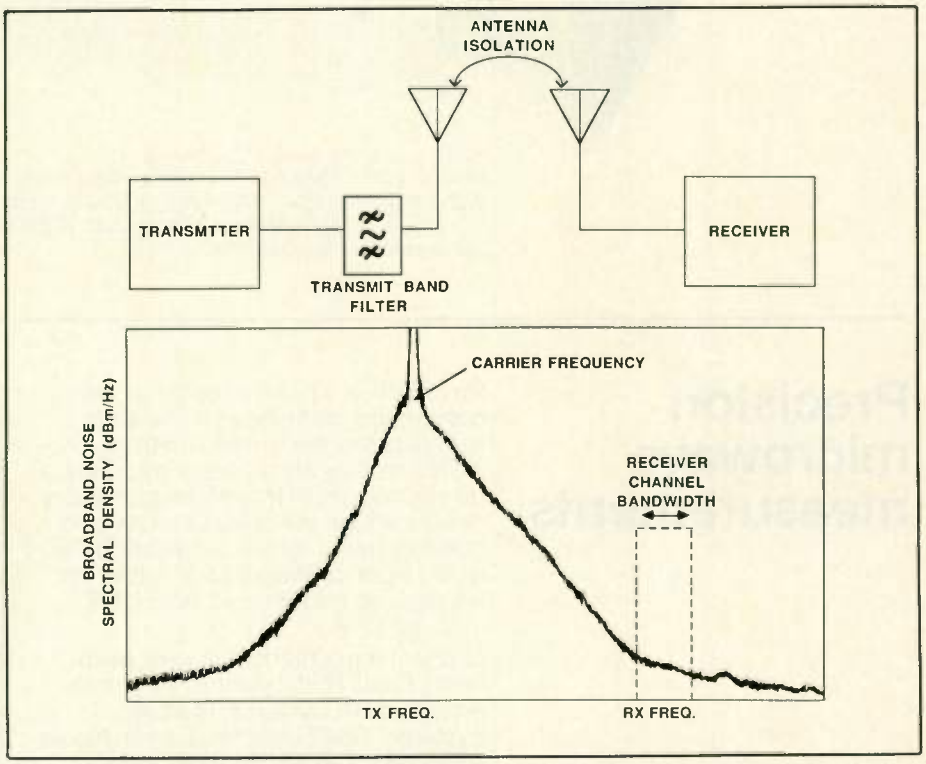


Figure 1. Broadband noise interference on duplex or co-located receivers.

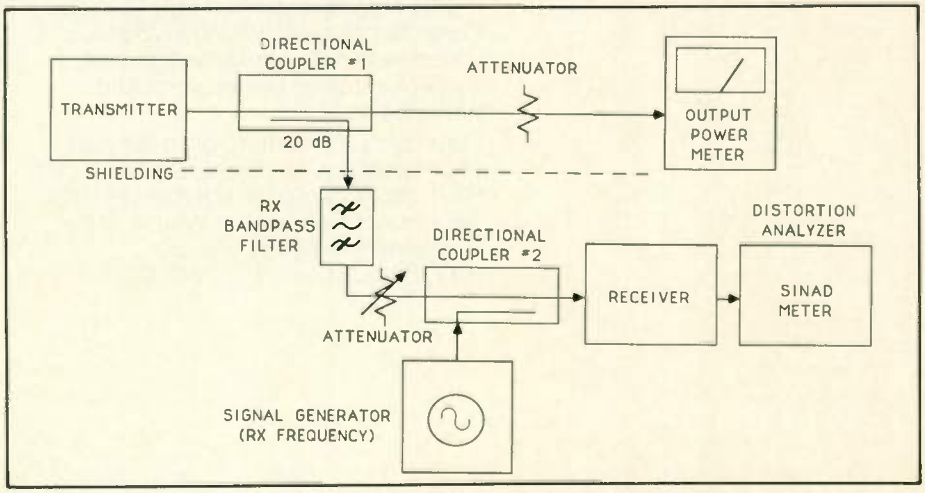


Figure 2. Transmitter noise interference test arrangement.

and power output are acceptable for correct operation, the remaining attenuation must be achieved by suitable transmitter filtering or antenna spacing to ensure that the level of broadband noise at the receiver input is also acceptable. The smaller the frequency separation and higher the transmitter power, the greater is the required filtering or antenna physical separation needed to achieve adequate isolation.

Imagine a typical transmitter carrier, Figure 1, a relatively clean carrier with no spurious signals. Broadband noise extends out beyond 30 MHz on from both sides of the carrier. Now imagine a receiver placed only 10 MHz away from the carrier. This is typical for duplex operation such as mobile radio telephone applications with several receivers co-located in a single site with many high transmitters. If the receiver and transmitter share the same antenna, the level of the noise signal present at the receiver channel without a duplexer is simply:

$$\text{Input Noise Power (dBm)} = \text{Noise Spectral Density (dBm/Hz)} + 10\log_{10}(BW_{IF}[\text{Hz}])$$

The total broadband noise is equal to the noise spectral density in dBm/Hz plus the integral of the broadband noise within the channel or IF bandwidth. The duplex filter (duplexer) must provide enough noise filtering of the transmitter signal to prevent receiver degradation. The co-located receiver situation differs from the duplex receiver problem only in the substitution of antenna isolation for the duplexer.

The measurement of the required attenuation of transmitter noise for duplex or co-location operation is shown in Figure 2. A signal generator is coupled to the receiver input and increased in amplitude until a 12 dB SINAD(1) (EIA Standard RS204) is achieved. With the transmitter energized, the attenuator is adjusted until the SINAD is degraded by 1 dB (11 dB SINAD). The sum of the loss of directional coupler #1 (20 dB), the receiver frequency bandpass filter, and the variable attenuator is the necessary isolation to produce only 1 dB reduction in SINAD. Shielding is very important in the measurement to prevent alternate paths of the transmitter noise from reaching the receiver. Attenuation values of 80 to 50 dB are required for frequency separations of 3 to 15 MHz for 100 watt transmitters. Typical required attenuation as a function of TX-RX frequency separation is shown in Figure 3. Exact curves are

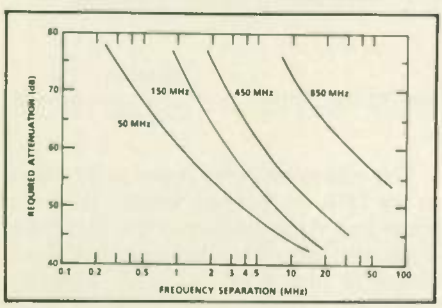
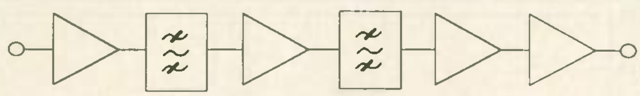


Figure 3.

GAIN/LOSS		+10dB	−2dB	+8dB	−2dB	+8dB	+8dB
NOISE FIGURE	±10MHz	17	–	20	–	23	25
	±30MHz	15	–	17	–	19	22
SELECTIVITY	±10MHz	–	20	–	20	–	–
	±30MHz	–	30	–	30	–	–



CARRIER POWER		0.15W	1.5W	0.95W	6.0W	3.8W	24W	150W
NOISE POWER (dBm/Hz)	±10MHz	-95.0	-85.0	-107.0	-99.0	-121.0	-113.0	-105.0
	±30MHz	-130.0	-120.0	-152.0	-142.8	-174.0	-146.9	-137.8

Figure 4. Power amplifier noise level block diagram.

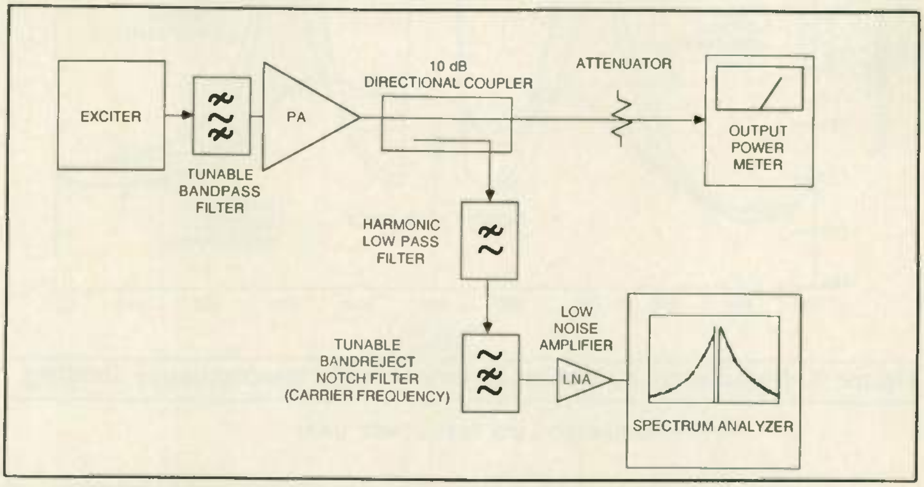


Figure 5. Broadband noise measurement of power amplifier.

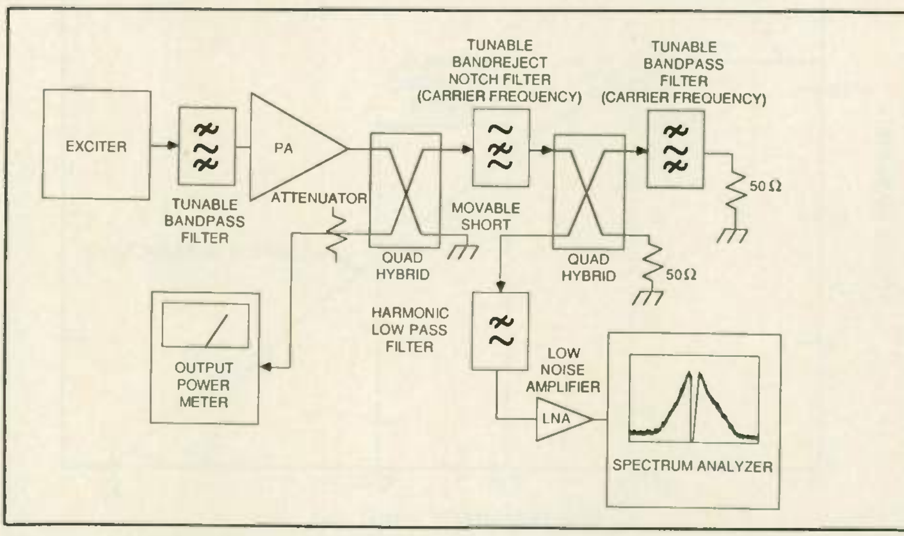


Figure 6. Alternate broadband noise measurement equipment layout.

constructed using the exact transmitter and receiver pair under construction.

Two methods are available for control of transmitter broadband noise interference. The first method is to provide sufficient isolation between the transmitter output and input to any receiver. This may be accomplished by physical separation of antennas or by increasing directivity. Transmitter filtering provides the second method of controlling interference from broadband noise. Filters may be added between amplifier stages or optimally placed at the transmitter output. A broadband noise power level diagram should be constructed (Figure 4), of the entire amplifier chain to determine the optimal effect of filtering. The maximum benefit gained from a filter occurs when the filter is placed as close to the output as possible. The noise figure is different at various offsets from the carrier frequency.

Frequency management (other than ensuring a wide spacing between the transmitter frequency and any receiver frequency) is only effective for narrow band predictable interference such as spurs and intermod products and not for broadband transmitter noise. Filters placed at the receivers will help to prevent receiver de-sensitization due to the presence of a large carrier signal but will not help the problem of broadband transmitter noise interference. The desired receive signal and the transmitter sideband noise are both at the same frequency.

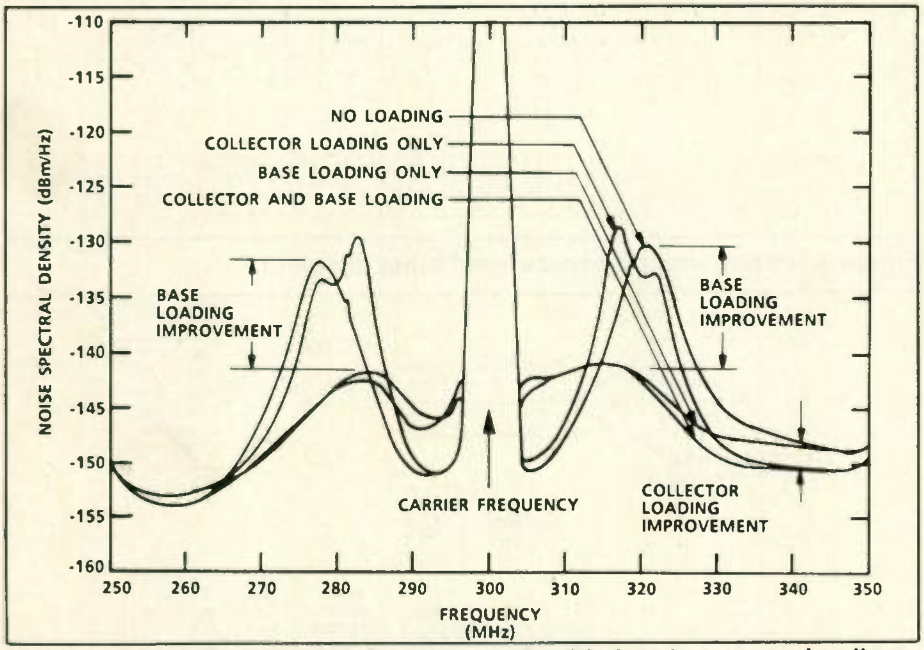


Figure 7. Broadband noise improvement with low-frequency loading.

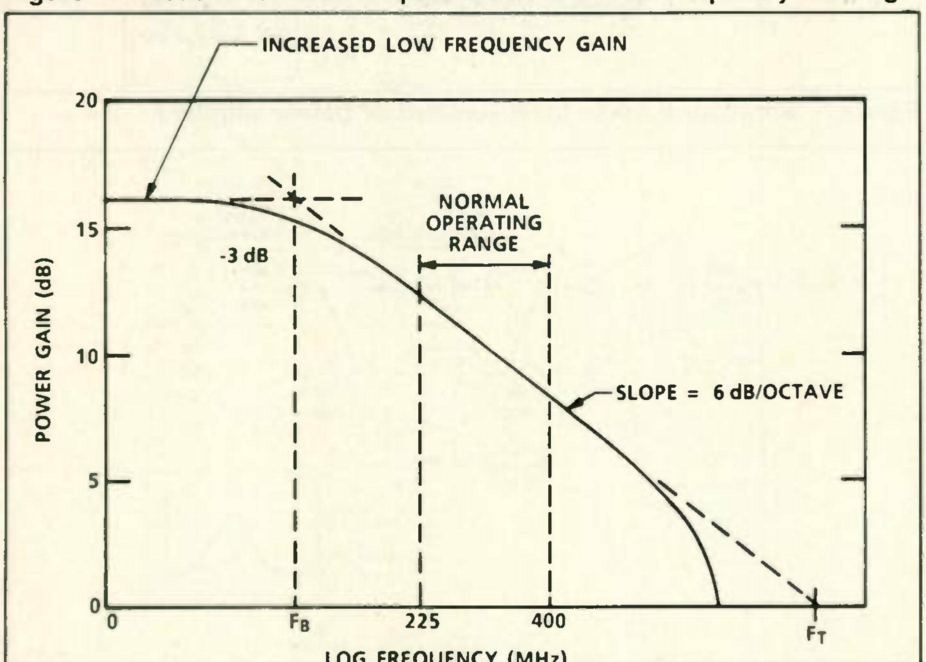


Figure 8. Gain slope for RF power transistors.

Broadband Noise Measurement

The sensitive measurement of broadband noise emissions from a power amplifier (Figure 5) requires nulling of the high level transmitter carrier to prevent overloading the spectrum analyzer. High quality spectrum analyzers have an average displayed noise level at approximately the −142 to −148 dBm/Hz level (Noise figure of 32 to 26 dB). This measurement floor is achieved only if the input RF attenuator of the analyzer is positioned at 0 dB. A low noise amplifier, LNA, (NF≤3 dB) with sufficient gain (≥25 dB) will set the overall noise figure (NF) of the measurement system according to Friis' noise equation:

$$NF = 10 \log_{10} F_1 + \frac{F_2-1}{G_1} + \frac{F_3-1}{G_1G_1} + \ldots$$

$$\ldots \frac{F_N-1}{G_1G_2 \ldots G_N}$$

where F_N=Noise *factor* of the Nth Stage

G_N = *Absolute* gain of the Nth Stage

This equation shows that the noise figure (10 log (noise factor)) of the measurement system (LNA/Spectrum Analyzer) is set by the low noise amplifier, if the gain is sufficiently high.

$$NF = 10 \log_{10} \left[10^{NF_1/10} + \frac{10^{NF_2/10} - 1}{10^{G_1/10}} \right]$$

$$NF = 10 \log_{10} \left[10^{3/10} + \frac{10^{26/10} - 1}{10^{25/10}} \right] = 5.1\text{dB}$$

The equivalent noise power at the input to the LNA must be subtracted from the measured noise power on the spectrum analyzer display. Thus the equivalent input noise at the LNA is −169 dBm/Hz (kTB+NF). A narrow band "clean-up" filter was placed at the power amplifier input to reduce the sideband noise to kTB in order to measure the amplifier chain's contribution to broadband noise.

The broadband noise measurement setup uses a 10 dB directional coupler in order to minimize insertion loss while still providing an acceptable return loss at the carrier frequency. A low pass filter should be included to avoid overloading of the LNA by second and third harmonic energy. A band-reject (notch) filter is also required to avoid overloading the LNA at the carrier frequency. A minimum notch depth of 50 dB is typically required to avoid compression of the LNA. The 3 dB bandwidth of this filter must be less than the offset noise being measured. Calibration of the measurement system is accomplished using a network analyzer placed between the through line of the 10 dB directional coupler input and the LNA output.

An alternate equipment (2) layout for measuring broadband noise is shown in Figure 6. The output power from the power amplifier under test is equally divided by the 90° equal-split directional coupler. Each output of the hybrid is terminated by a short at the carrier frequency. The reflected power adds 180° out-of-phase at the hybrid input, to provide a good return loss at the carrier frequency, and in-phase at the power meter terminal, to provide a measurement of carrier power. Broadband noise offset from the carrier frequency undergoes 3 dB loss before passing through the notch filter. The broadband noise, offset from the carrier frequency, will be reflected and suffer a total of only 6 dB loss before appearing at the low pass filter.

It must be mentioned that the noise figure of a power amplifier stage should not be measured in the same manner as that of small signal amplifiers. The output broadband noise of a power transistor biased Class A is much lower than the value measured in the presence of a large RF voltage swing with the same collector current. "Class A" small-signal values of 15 dB noise figure, measured at an equivalent bias current, turn out to be on the order of 25 dB when operated Class C at rated output power.

Noise Improvement

The broadband noise performance of a bipolar transistor power amplifier is improved by the use of low frequency resistive loading. The measured results of this broadband noise improvement technique are shown in Figure 7, with a graph of broadband spectral noise density plotted on both sides of a carrier frequency. The presence of this close-in broadband noise power is not unexpected after considering

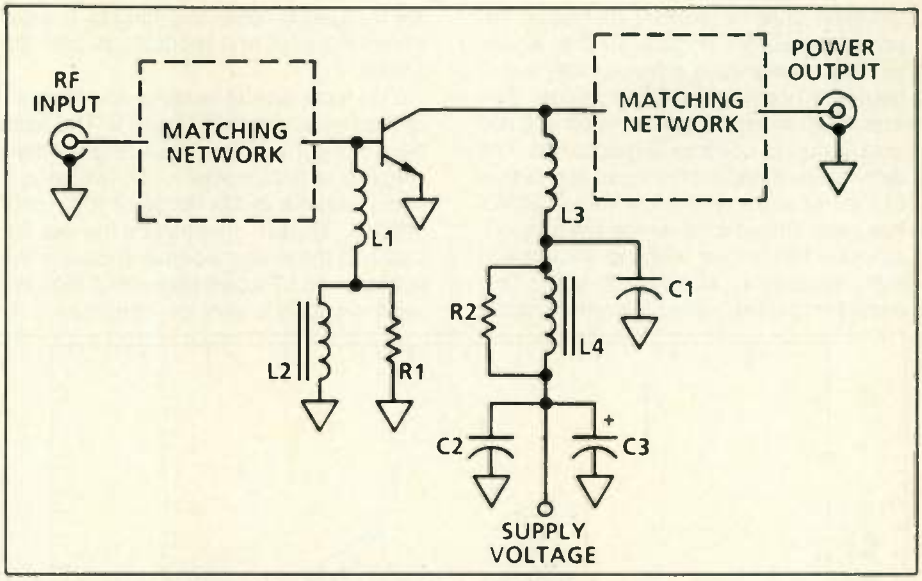


Figure 9. Collector and base loading circuits.

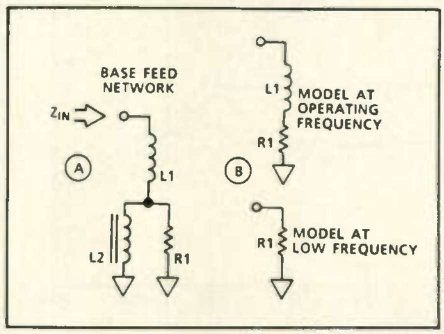


Figure 10. Base feed network modeling.

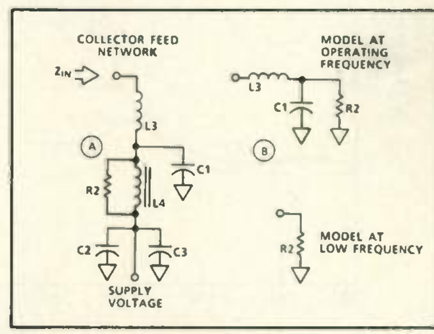


Figure 11. Collector feed network modeling.

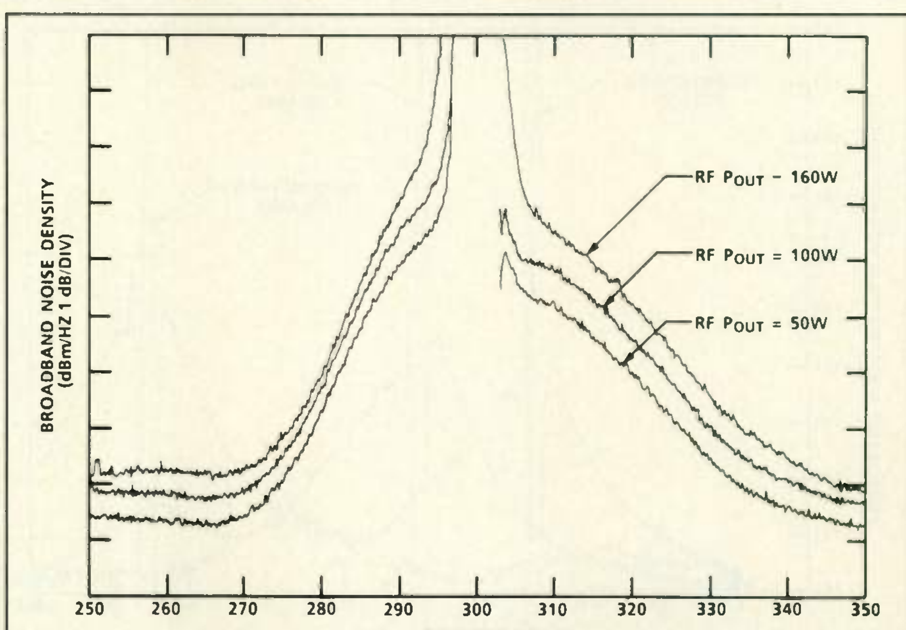


Figure 12. Broadband noise performance as function of output power.

the gain shaping present in bipolar RF power transistors (Figure 8). The device exhibits greater gain at frequencies lower than the normal operating region. The maximum available gain rolls off at 6 dB/octave due to collector capacitance. The technique of loading the input and output of the transistor with a low value resistor has been shown to lower the low frequency (≤50 MHz) gain with no sacrifice in high frequency (≥200 MHz) gain. This broadband noise power is predominantly low frequency noise amplified by the low frequency gain and modulated onto the carrier.

This technique of resistive low frequency loading is shown in Figure 9. The base bias current return circuit (Figure 10A) consists of two chokes — L1, which is a small inductor in the range of 100 nanohenries, chosen to function as an RF choke at the lowest operating carrier frequency, and L2 which presents a high impedance down to very low frequencies. At normal operating frequencies the base is effectively isolated by inductor L1 from resistor R1, forming a low Q arrangement (Figure 10B). The Q of the network is given as $Q = 2\pi F\ L1/R1$, which is low. The resistor R1 is chosen to be in the range of 5 to 10 ohms. At low frequencies, where the power gain is much greater, the shunting effect of R1 increases, due to the low inductive reactance of L1. The larger value inductor L2, consisting of a few turns of wire through a ferrite bead, is added to provide high impedance at low frequencies. Thus the base is effectively resistively loaded by R1 at low frequencies where the power gain is greater.

The collector feed network (Figure 11A) accomplishes a similar feat by resistively loading the collector at frequencies below the operating band. Within the mid-frequency range between the normal operating frequencies and the low frequency region, the small inductor L3 and the small shunt capacitor C1 act as an L-network to transform the resistance of R2 to a value that tends to resistively load the collector. At the operating frequency, the ferrite loaded inductor L4 appears as an open, while the capacitor C2 acts as a very low impedance to yield the collector feed network model of Figure 11B. Below the normal operating range, the input impedance to the network looks resistive, asymptotically approaching the value of R2, typically 10 to 20 ohms. The wattage of this loading resistance need only be ½ to 1 watt because it is designed to absorb less than ¼ watt of signal power at the operating frequency. The bypass capacitors, C2 and C3, placed at the supply voltage, consist of a disc ceramic for bypassing at RF and an electrolytic for bypassing at audio frequencies.

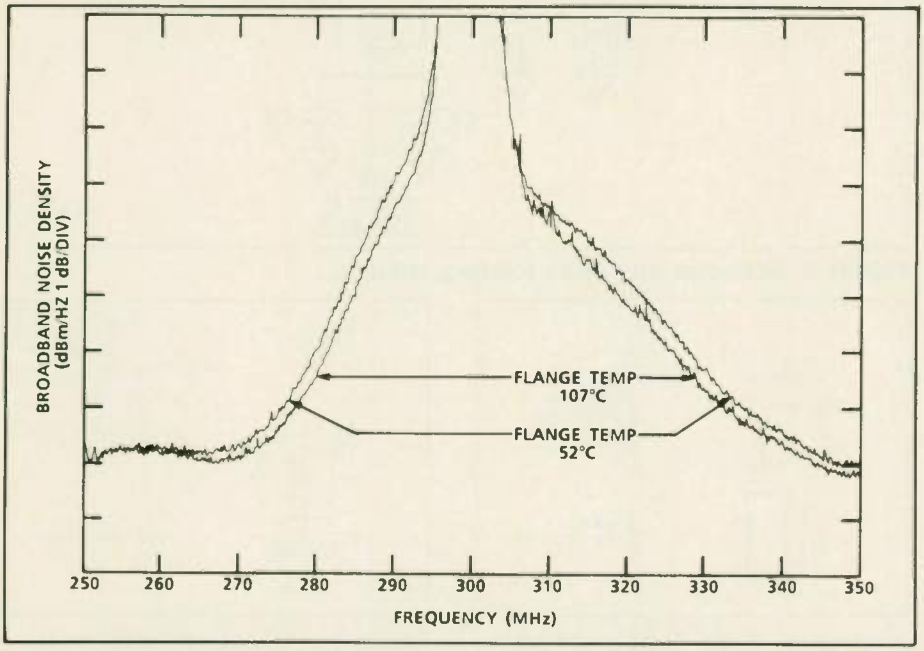


Figure 13. Broadband noise performance at elevated temperatures.

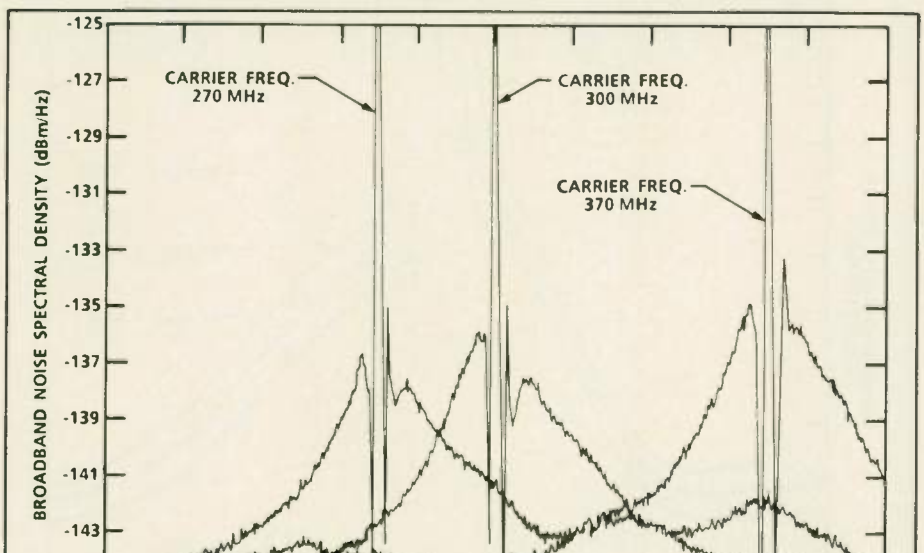


Figure 14. Wideband spectral shape of broadband noise.

Variations in Broadband Noise

Variations in broadband noise power as a function of RF output power, flange or junction temperature, frequency offset, and circuit element values have all been considered. Broadband noise power is shown as a function of output power from a 150 watt UHF power amplifier in Figure 12. As the RF output power increased from 50 watts to 160 watts (5 dB increase), the broadband noise power increased by only 1 dB. Thus, broadband noise is not a strong function of the actual level of output power and appears to be a collector-base breakdown phenomenon starting at a power level of about 1/10 rated output.

Broadband noise was also examined as a function of flange temperature (Figure 13). Broadband noise power remained relatively constant when the flange temperature of the RF power transistor increased from 52°C to over 100°C (junction temperature = 150°C). Next, the level of broadband noise was examined as a function of frequency offset from the carrier frequency. The shape of the noise (Figure 14) is quite smooth and monotonically decreases beyond the ±30 MHz offset.

Finally the effects of variations of the resistive-loading circuit element values were considered. Variations in the value of the base loading resistor are shown in Figure 15, and the collector loading resistor, Figure 16. Expected variations in the chosen value of the base or collector resistor result in less than 1 dB uncertainty in broadband noise power at ±30 MHz offset from the carrier. Low frequency resistive-loading also improves amplifier (3,4) stability for operation into various mismatched load impedances.

FET

The Field Effect Transistor has been shown to produce (5-8) typically 10 to 15 dB lower broadband noise than a comparable bipolar power transistor. The voltage avalanche breakdown characteristic of the bipolar is not present in the FET. Care must be exercised, however, to avoid electro-static discharge of the gate which may ruin the noise performance of the device. Devices are presently capable of producing RF output power in excess of 50 watts for operation both in the VHF and UHF frequency range. It is anticipated that within the next few years that FETs will have fully matured to the 150 watt power level through the UHF frequency band.

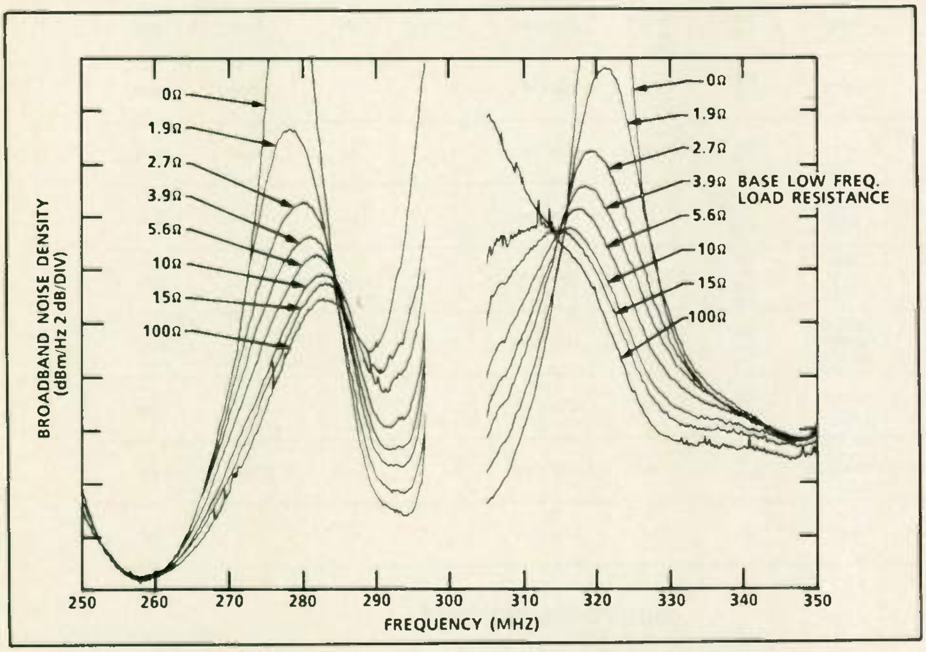


Figure 15. Variations of the resistive load of base circuit only.

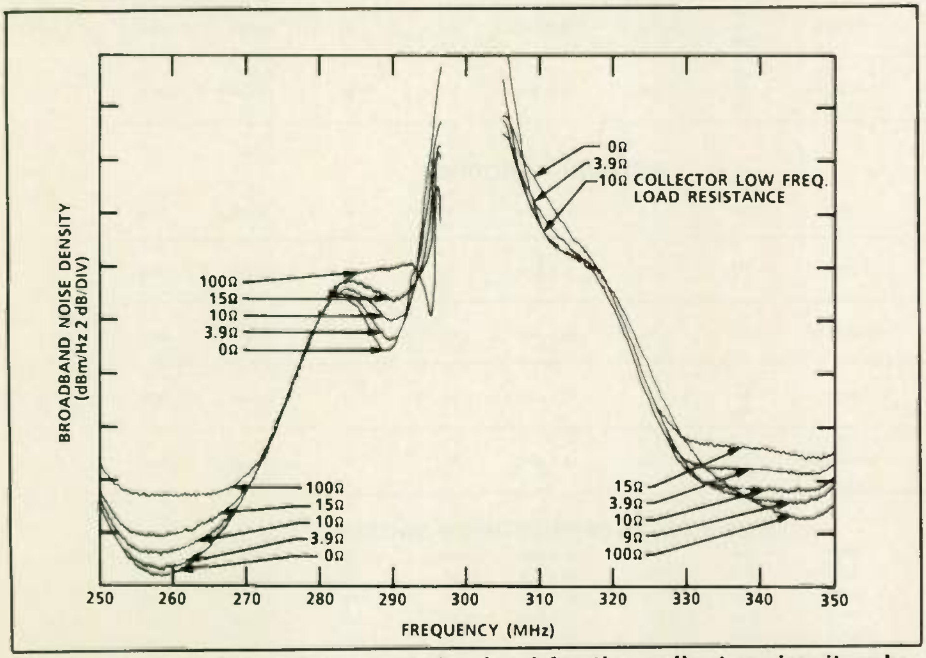


Figure 16. Variations of the resistive load for the collector circuit only.

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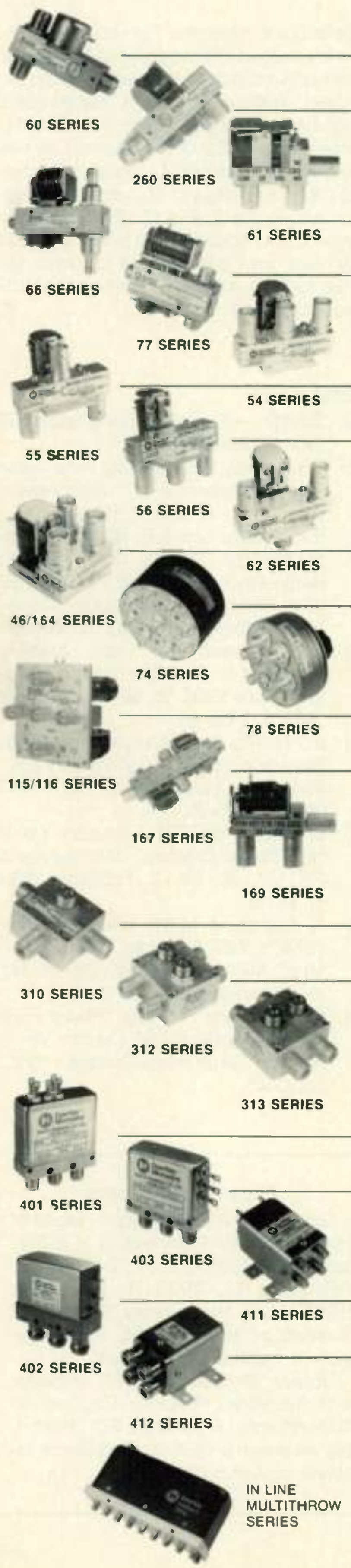

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61	SPDT	BNC	50	2000 MHz	0.1	50	150W	50W
66	SPDT	F	75	1000 MHz	0.1	70	100W	20W
77	SPDT	F BNC	75 50	2000 MHz	0.1	40	100W	20W
54	SPDT	BNC TNC	50	4000 MHz	0.1	70	100W	50W
55	SPDT	BNC TNC	50	4000 MHz	0.1	70	100W	50W
56	SPDT	BNC TNC	50	4000 MHz	0.1	70	100W	50W
62	SPDT	BNC TNC	50	4000 MHz	0.1	70	100W	50W
46 164	DPDT SPDT	BNC SMB SMA	50	4000 MHz	0.1	50	125W	50W
74	SP6T SP7T SP8T	BNC N TNC SMA	50	2500 MHz	0.1	50	100W	20W
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115 116	SP3T SP4T	BNC SMA	50	3500 MHz	0.1	50	200W	50W
167	TRANSFER	N UHF	50	1000 MHz	0.1	40	1000W	100W
169	SPDT	BNC	50	1000 MHz	0.1	50	100W	20W
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312	BYPASS	SC N HN	50	1000 MHz	0.1	30	2000W	500W
313	TRANSFER	SC N HN	50	1000 MHz	0.1	30	2000W	500W
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411	TRANSFER	SMA	50	26.5 GHz	.1 TO .5	70	200W	25W
402	SPDT	BNC N TNC	50	16.0 GHz	.1 TO .5	70	1000W	100W
412	TRANSFER	BNC N TNC	50	16.0 GHz	.1 TO .5	70	1000W	100W
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473	SP7T	SMA	50	18.0 GHz	0.7	60	100W	10W
483	SP8T	SMA	50	18.0 GHz	0.8	60	100W	10W
493	SP9T	SMA	50	18.0 GHz	1.0	60	100W	10W
4A3	SP10T	SMA	50	18.0 GHz	1.0	60	100W	10W
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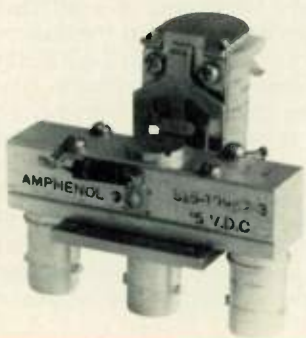

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PIN Diode Switches — Part I

Some Notes on the Characteristics and Performance of Shunt Reflective Type PIN Diode Switches.

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

In this application a PIN diode behaves like a simple mechanical SPST switch. To stop the flow of RF energy it provides a short-circuit across the RF path. Thus, most of the RF energy is reflected rather than absorbed by the switch. However, a PIN diode is an imperfect switch, and its imperfections have to be taken into consideration.

In this switch's "off" condition (diode conducting), the most important diode characteristic to consider is the series diode resistance (R_d), Figure 1(b). The isolation (switch OFF-diode ON) increases with a decrease in R_d, as shown in Figure 2. Putting diodes in parallel improves isolation by decreasing the total R_d. The effect is the same as using a single diode with an effective R_d equivalent to all the diode resistances in parallel. Thus, the isolation improvement is 6 dB each time the number of diodes is doubled, a somewhat inefficient method. Fortunately, performance can be considerably improved using comparatively simple methods.

A section of a transmission line placed between any two diodes produces a large improvement. If the line impedance is the same as the load impedance, its presence will not affect the switch performance in the ON condition. A two diode switch, as shown in Figure 3, will be used to illustrate this effect. However, this analysis can be extended to multi-diode arrangements. The effect of the transmission line is shown in Figure 4. The transmission line transforms the second diode ON resistance, R_{d2}, to a high equivalent series value at the location of the first diode, thus increasing the effect of the first diode. Or, conversely, it increases the effective source impedance, seen at the second diode location, as shown in Figure 3(b). As in all cases when transmission lines are used for transforming impedances, the effect is frequency sensitive. Fortunately, in this case, this effect is quite broadband, as shown in Figure 4. The effect is maximum when the line is 90 degrees at the operating frequency. When line length is 0 or 180 degrees, the effect is the same as placing the diodes in parallel.

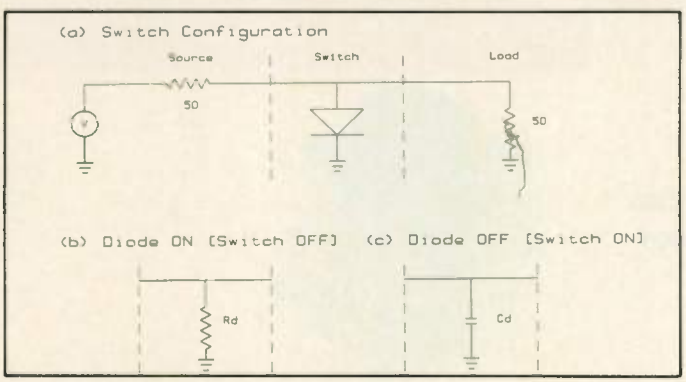


Figure 1. Shunt PIN diode switch

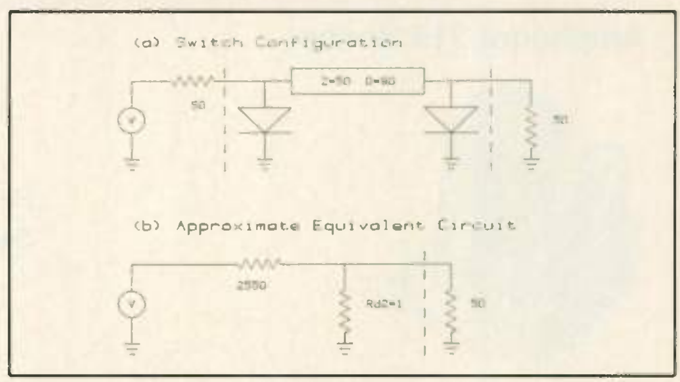


Figure 3. 2-PIN diode switch.

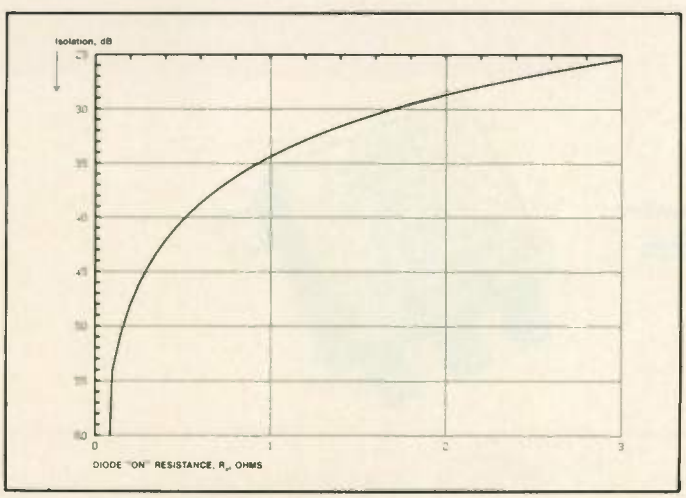


Figure 2. Switch isolation

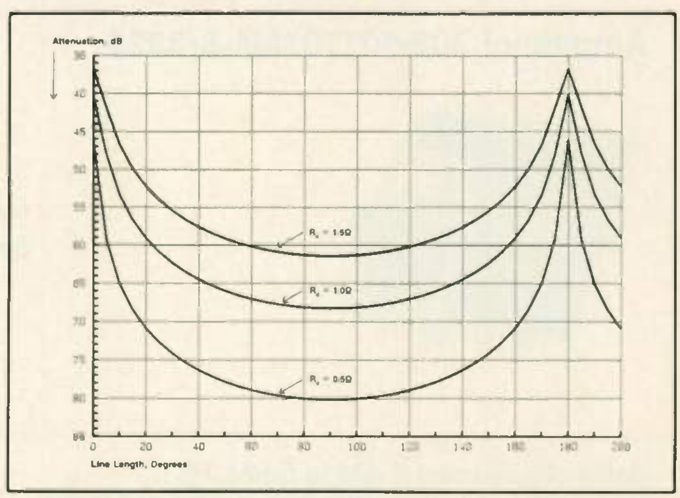
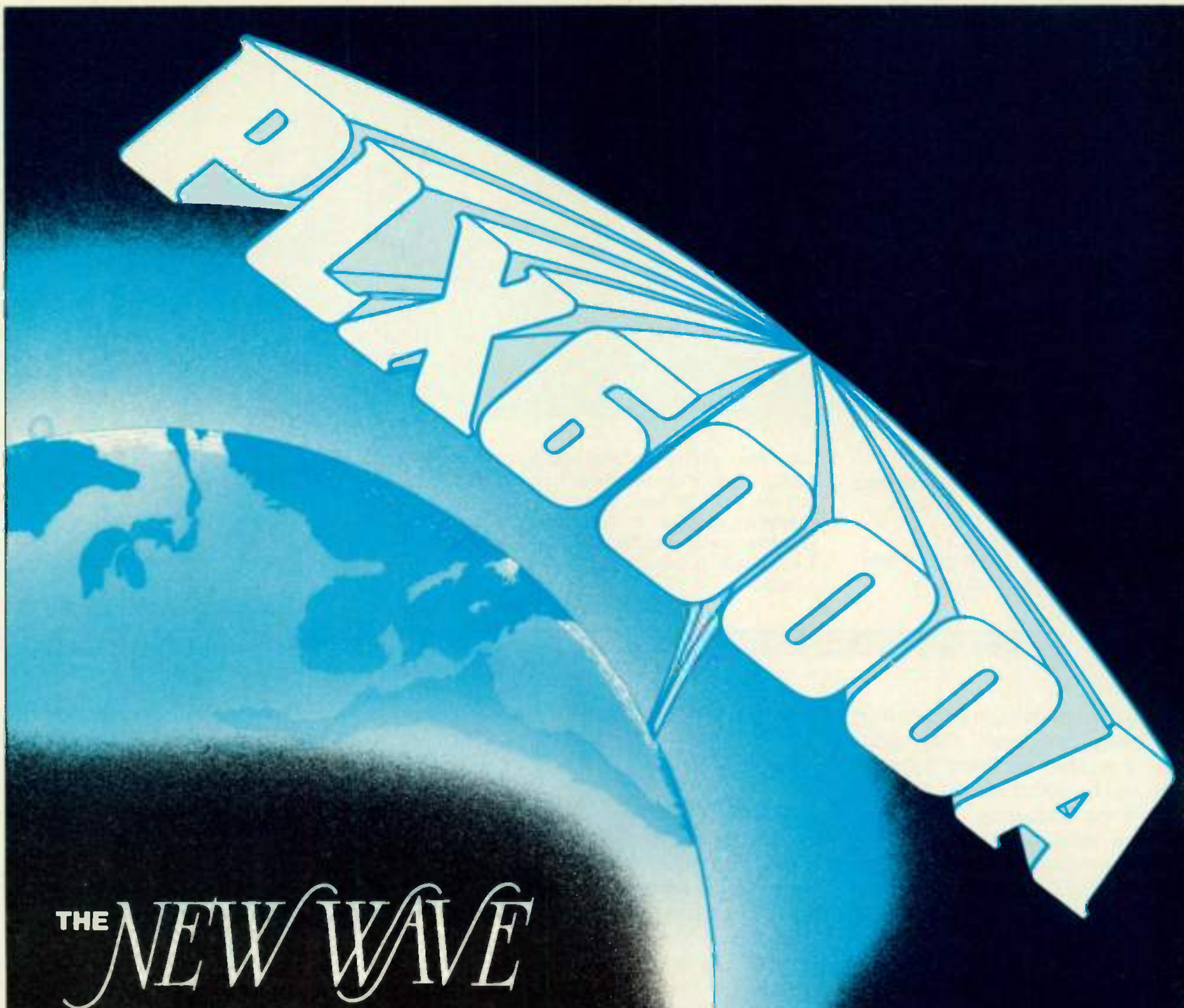


Figure 4. Isolation of 2 diode switch.



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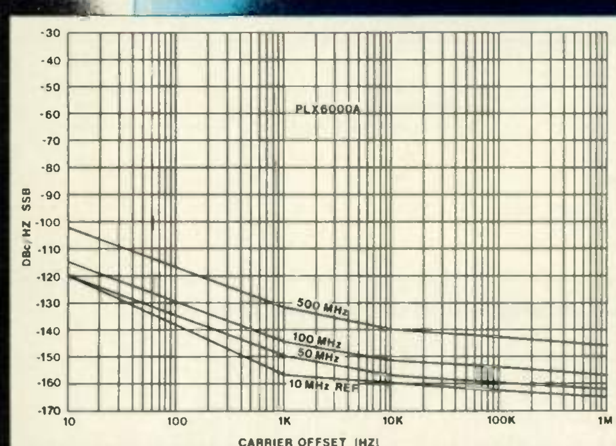

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Thus, to obtain maximum benefit from multiple shunt-configured PIN diode switches, the line sections between diodes should be about 90 degrees long at the operating frequency.

It should be noted that, so far, a 50 ohm source was assumed. An impedance source lower than 50 ohms will deteriorate the isolation. This can be remedied, however, by placing another 90 degree line between the source and the first PIN diode, as shown in Figure 5. Figure 6 shows the effect of the source impedance on isolation and Figure 7 shows the effect of the source impedance on isolation when the input transmission line is introduced.

Shunt Switch — "ON" Condition

In the switch ON condition, the PIN diode is back-biased and presents a high resistance across the transmission line, which usually can be neglected. The important diode parameter becomes the diode capacitance, C_d, shown in Figure 1(c). This spurious reactance introduces some additional losses. This unwanted insertion loss, in the ON condition, can be "tuned out" using inductors in parallel with the diodes. This compensation is frequency dependent, but, since the Q of the tuned circuit is very low, the compensation is effective over a large bandwidth. It is also possible to compensate for this reactance by changing the line length between the diodes. This effect, using Figure 3(a) circuit (C_d = 1 pF and an operating frequency of 2000 MHz), is shown in Figure 8. As usual, therefore, some compromises exist and have to be considered.

Part II of these PIN diode notes will review the series configuration of switches. Look for it in an upcoming issue.

About the Author

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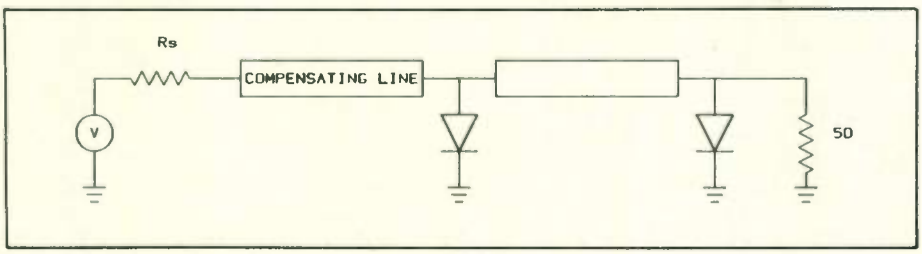

Figure 5. Compensation using input line.

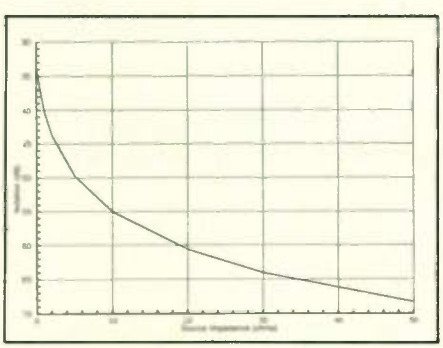

Figure 6. Effect of source impedance (R_{d1} = R_{d2} = 1 ohm).

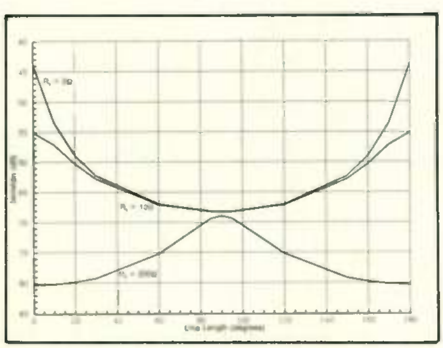

Figure 7. Effect of input transmission line.

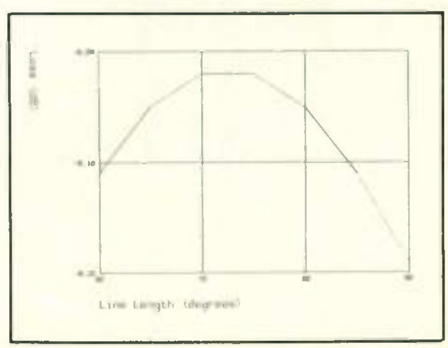

Figure 8. Compensation with line length.

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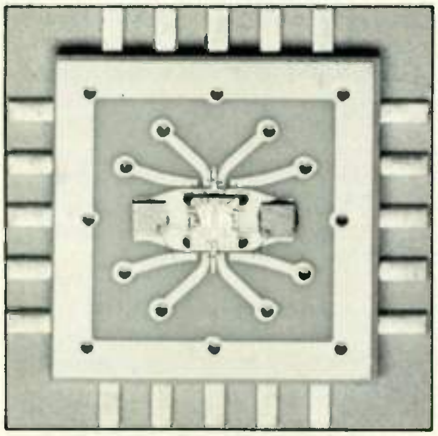
Broadband GaAs Microwave Interface Chips

TriQuint Semiconductor announces a new line of GaAs integrated circuits designed to simplify the construction of microwave systems operating in the 1 to 8 GHz frequency range. The MICRO-S series includes the TQ9111 7 dB amplifier, the TQ9141 two-way power splitter, the TQ9151 SPDT switch, and the TQ9161 electronically variable attenuator. They cover 1 to 8 GHz with uniform amplitude



characteristics (variation less than ±1 dB, typically). Prices for the MICRO-S microwave component series are as follows. In die form for quantity 100, the TQ9111 is $60, the TQ9141 is $70, the TQ9151 is $51, and the TQ9161 is $70. Corresponding prices for the 18 GHz surface mount package are: TQ9111 ($114), TQ9141 ($155), TQ9151 ($143) and TQ9161 ($155). **TriQuint Semiconductor, Inc., Beaverton, Ore. INFO/CARD #130.**

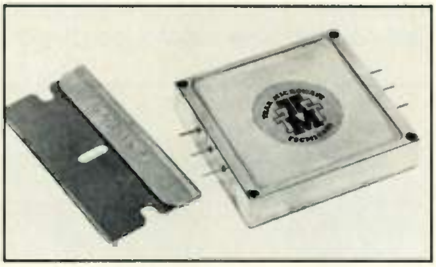
EMI Analyzer/Receiver Covers 0.5-40 GHz

Electro-Metrics announces the latest addition to its line of microprocessor-controlled electromagnetic interference (EMI) analyzer/receivers, the EMC-60 MK IV. The new unit performs emission and susceptibility testing to strict military, government and commercial specifications, including MIL-STD-461/462 and FCC Part 18, from 0.5 to 40 GHz. In addi-

tion, the unit's front end can be remotely operated and controlled to reduce input cable losses and increase flexibility. Advanced detector circuits provide true average, direct peak and slideback peak analysis. **Electro-Metrics, Amsterdam, N.Y. INFO/CARD #129.**

Crystal Oscillator is Drop-in Packaged

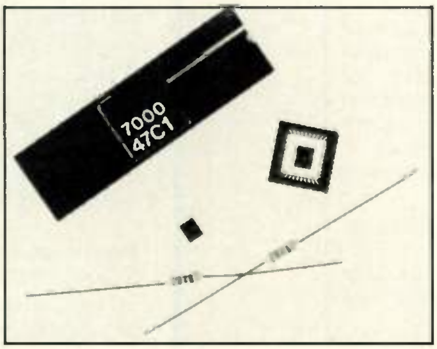
TRAK Microwave introduces a new drop-in crystal controlled oscillator intended for use in surface mount/microstrip designs. The unit is only 1.33 × 1.25 × .28 inches and weighs only .44 ounces.



Frequency range is 60-1200 MHz, with a power output of 10 mW or 100 mW. One watt versions are available over limited temperature range. **TRAK Microwave Corporation, Tampa, Fla. Please circle INFO/CARD #128.**

Logarithmic Video Amplifier Operates up to 30 MHz

A universal logarithmic amplifier integrated circuit, the M-7000, is designed to work over a 30 MHz wide bandwidth. The M-7000 is designed as seven precision modular logging stages. The logging



stages are independent of one another and the designer may use each stage separately or cascade any number together. Each stage has a user defined dynamic range of 6 to 17 dB. The typical RMS conformity of the output is .5 dB for a 105 dB dynamic range. The user may customize the transfer response. The IC contains a differential video summing amplifier, a band gap reference, a precision die temperature sensor and a dual tracking voltage regulator. Price is $27 for quantity 1 to 100 commercial, plastic 40 pin dip and $126 up to quantity 100, Mil grade. **Megadyne Corporation, Fairfax, Va. INFO/CARD #126.**

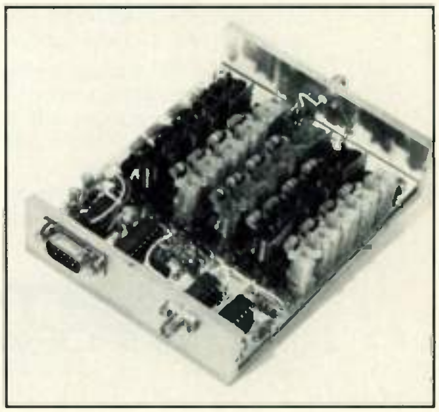
Miniature Quartz Crystals

Chemically milled miniature AT quartz crystals for Pierce, Colpitts and Series oscillators are available in rugged, hermetically-sealed miniature ceramic packages (CX) with standard leads or with leadless contacts for surface mounting by reflow solder, conductive epoxy or other techniques. These CX-AT quartz crystals

are available in the 10 to 24 MHz frequency range. Dimensions of the CX leadless package are: 8.38 mm (.33") × 3.94 mm (.155") × 2.03 mm (.08"). **ETA Industries, Inc., New York, N.Y. Please circle INFO/CARD #127.**

Switched RF Filter Has Five Sections

Interad Ltd. announces a switched RF filter bank Model 2425A, a single module consisting of five filters of five poles each under the control of a three-bit code. The passband frequency range of this model is 56 to 74 MHz, but is is available in center frequencies of 1-100 MHz and bandwidths of 3 percent to 20 percent



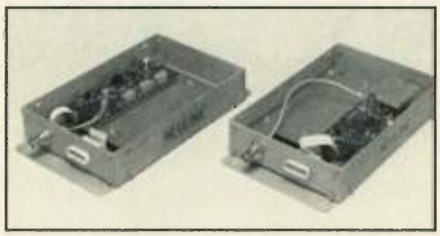
F_c. The Model 2425A has the following major frequencies: passband center frequency range of 56 to 74 MHz, 3 percent to 10 percent bandwidth, 5 dB max. insertion loss, and VSWR input of 1.5:1 max. Price of the unit is $675. **Interad Ltd., Gaithersburg, Md. INFO/CARD #125.**

Vacuum Relays Switch from DC to 50 MHz

Three series of vacuum relays are now available from Siemens Components, Inc. Designated the "Series 100," "Series 300," and "Series 400," all units offer high-voltage, high current switching capability, as well as the ability to switch extremely small signals from DC to 50 MHz. Applications include RF transmission, to control RF circuits without load switching, where extremely low transfer resistance and high insulation resistance are important; and load switching, to control loads of several kilowatts at DC or 60 Hz voltages, with the ability to keep the discharge low. **Siemens Components, Inc., Iselin, N.J. Please circle INFO/CARD #124.**

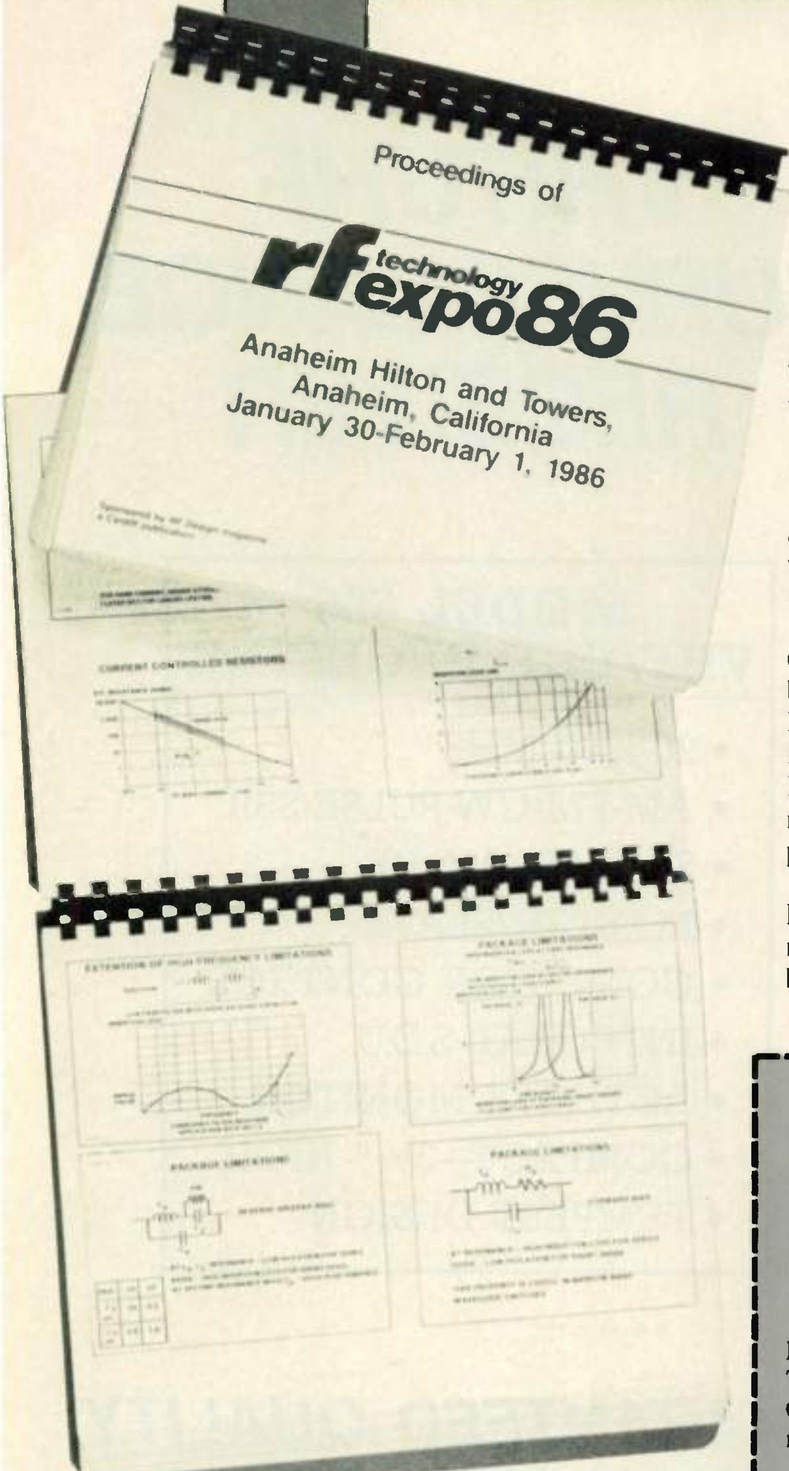
900 MHz RF Link Transfers Voice or Data

Neulink introduces a 928-960 MHz RF link for speech communications applications. The model RFL-T9S transmitter and RFL-R9S receiver are designed to transfer voice and tone communications data. Applications include control linking of repeaters and base stations operating on various frequency bands. The RFL-T9S transmitter has adjustable output power from 1-2 watts, and is continuously rated. Both units come housed in separate aluminum alloy enclosures, 9" × 5" × 1½". **Neulink, San Diego, Calif. Please circle INFO/CARD #123.**



GaAs Digital Components are Under $100

GigaBit Logic has announced the availability of three new PicoLogic™ family ICs all priced below $100. The new gallium arsenide (GaAs) ICs include an octal register/shift register, dual 9-bit parity generator and checker/8-bit word comparator, and 4-bit synchronous cascadable counter. All three new devices are ECL I/O and 10G PicoLogic™ I/O compatible. A V_{bb} reference input allows threshold tracking when interfacing to ECL logic families. The output rise and fall times are typically 150ps. In a leadless chip carrier, or leaded chip carrier, the prices (100s) are: 10G022 ($79.50), 10G045 ($46.00), and 10G061 ($99.50). Prices are lower for dice form. **GigaBit Logic, Newbury Park, Calif. Please circle INFO/CARD #121.**



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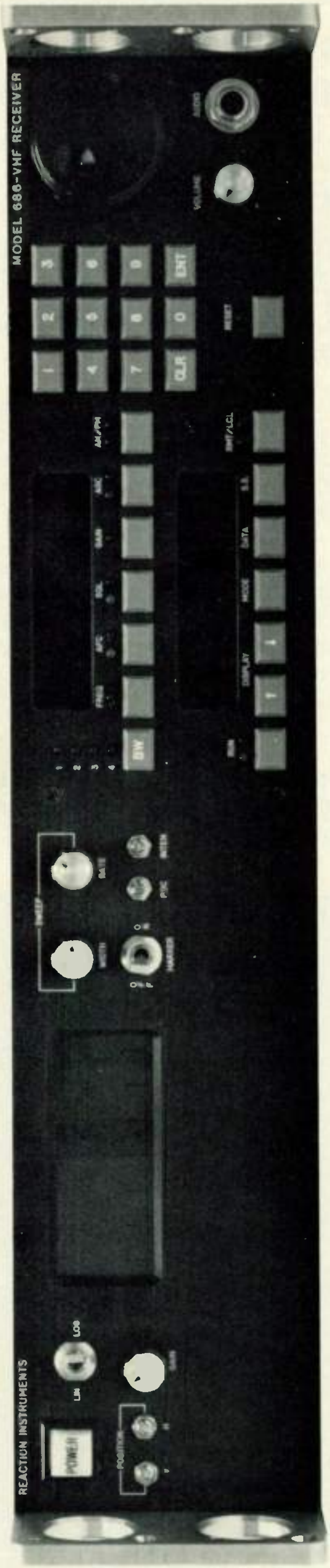
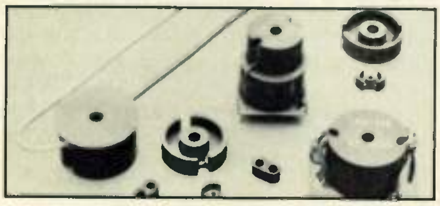

MODEL 686-VHF RECEIVER
REACTION INSTRUMENTS
POWER

Ferrites and Wound Components are Introduced

Iskra Electronics has announced a complete line of ferrites, cores and wound components for filters, transformers, RFI suppression and other industrial and commercial applications. Iskra will be producing a wide variety of ferrite core types and configurations, made of several different material grades. The Iskra line will include Pot E, RM, toroid and rod bead



cores, X cores (with gaps and adjustments). Iskra wound components consist of stable ferrite cores with high temperature and time stability, permeability and small losses. The components are not standardized, therefore all required characteristics are customer designed. **Iskra Electronics, Inc., Farmingdale, N.Y. INFO/CARD #122.**

High Power Attenuator Handles 50 Watts

RLC Electronics announces its Model A501, a 50 watt high power fixed attenuator. This attenuator features excellent power handling capabilities, low VSWR and extreme accuracy under all environmental conditions. This unit has a thick film element for stability and long life while operating over a frequency range of DC to 5 GHz. The unit comes in values of 3 dB, 6 dB, 10 dB, and 20 dB with an attenuation accuracy of ±.5 dB over the specified frequency range. These units are available with Type "N" and "SMA" Type connectors. Prices start at $275 in unit quantities. **RLC Electronics, Inc., Mt. Kisco, N.Y. INFO/CARD #120.**

Dual Directional Coupler Covers TACAN Band

Sage Laboratories is offering a new TACAN band dual directional coupler. The Model FC3364, which is currently being delivered for use on a new FAA system, handles 10 kW peak, 100 watt average input power at 1030 MHz. Coupling is flat over the 1010 to 1110 MHz range. The forward coupling is 20 dB and the reverse coupling is 15 dB. Directivity is 35 dB minimum for the reverse coupler. Net main line insertion loss is less than 0.25 dB. The couplers are also available as phase and amplitude tracking parts. The part number for the matched pair is FC3364-1. **Sage Laboratories, Inc., Natick, Mass. INFO/CARD #119.**

Miniature Frequency Standard Has High Stability

Piezo Systems announces the availability of a miniature, high stability frequency standard. The Piezo Systems Model 2810005 combines the latest techniques in crystal design with a low noise proportional oven control circuit. The oscillator has an aging rate of 5×10^{-8} per day. The unit obtains final frequency within 1×10^{-7} after five minutes at 25°C. This model is available in 1.0 MHz to 20.0 MHz range. Frequency stability is $\pm 5 \times 10^{-8}$ for any 10°C increment over 0°C to +70°C. **Piezo Systems, Carlisle, Pa. INFO/CARD #118.**

RF Hybrid Amplifier for 20-400 MHz

Amplifonix has announced a new RF hybrid amplifier, Model TM 6441 in a TO-8 hermetic package, with gain of 14.5 dB typical over the frequency range of 20 to 400 MHz. Other specifications include: noise figure, 3.8 dB typ.; power out @ 1 dB comp., +16 typ.; VSWR (in/out) of 2:1; current @15 V, 32 ma.; temperature range of −55 to +85°C. All units meet MIL-STD-883B screening. Price for 1-9 is $84.00. **Amplifonix, Inc., West Langhorne, Pa. INFO/CARD #117.**

Oscillator Syncs to External Trigger

Berkeley Nucleonics Corp. has introduced a compact new type frequency source known as the Model C-1100 trigger-coherent oscillator. It is available with one, or optionally two, selected outputs between 850 kHz and 115 MHz, supplied in ECL pulse train form. The C-1100 acts as a fixed-frequency clock-pulse source that repeatedly syncs its output to unpredictable external events represented as signals to the unit. The unit can be steadily or randomly triggered at 0 to 20 MHz rates. Applications include any situation where clock phasing to outside events is important. Price of the C-1100 ranges from $280 to $350. **Berkeley Nucleonics Corp., Berkeley, Calif. INFO/CARD #116.**

FET Amplifiers Have Wide Dynamic Range

TRW Microwave has announced a new series of 2 to 6 GHz FET amplifiers. These miniature IF amplifiers feature high intercept and low noise figure, and are ideal for wide dynamic range EW system applications. They provide +27 dBm minimum power output at 1 dB gain compression, and a variety of small signal gains. Temperature compensation from −54°C to +95°C is provided using a unique PIN circuit. Gain flatness over frequency and temperature is ±1.5 dB, maximum. Noise figure, with a nominal gain of 40 dB, is 3.5 dB maximum at +25°C and 4.5 dB maximum at +95°C. Package size is 0.32" × 0.66" × 2.50". **TRW Microwave, Sunnyvale, Calif. INFO/CARD #115.**

RF Signal Sampler Simplifies Measurement and Control

RF Industries announces the introduction of its new RFA-4059 RF signal sampler. It is made of silver plated machined brass with gold center pins and PTFE insulation throughout. The unit comprises an "N" female connected to an "N" male by a thru line section. The vertical part of the T is terminated by a BNC female jack, from which the sampled RF is taken. This BNC section of the signal sampler is movable, and can be inserted or withdrawn from the sampler to either increase or decrease the amount of signal that is sampled. Attenuation from 20 to 80 dB (depending on frequency) can be produced by just adjusting the BNC

rf products *Continued*

slug. The signal sampler can be used at frequencies from 50 MHz-12 GHz, and at powers of up to 500 watts continuous, 5 kW peak. **RF Industries, Hialeah, Fla. INFO/CARD #113.**

CMOS Digital Multipliers Reach 30 MHz

Analog Devices has introduced four new 8-bit and 12-bit digital multipliers and multiplier/accumulators (MACs) in its high-speed 1.5-micron CMOS process. These four devices are significantly faster than the bipolar parts they replace, yet they consume less than 10 percent of their power. Applications for 8- and 12-bit multipliers and MACs include video, imaging, radar, and telecommunications. In 100s, plastic DIPs range from $27.00 to $53.30. **Analog Devices, Inc., Norwood, Mass. INFO/CARD #114.**

Crystal Oscillators Lockable to 5 MHz or 10 MHz

Communications Techniques, Inc. has announced their Series PXS, single frequency crystal oscillator, phase-locked to 5 or 10 MHz standard, with 30 to 140 MHz frequency range and low phase-noise. The internal crystal oscillator is phase lockable by digital synthesis techniques to an external frequency standard at either 5 to 10 MHz for improved close-in (< 100 Hz) phase-noise as well as improved long-term frequency noise performance (depending on the reference standard used). The output frequency does not have to be an integer multiple of the reference frequency. The digital synthesis technique used enables the output frequency to be set to almost any desired frequency. **Communications Techniques, Inc., Whippany, N.J. INFO/CARD #112.**

TV Stereo Test Generator Has 60 dB Separation

RE Instruments Corporation introduces the Model RE540 TV Stereo Generator, providing accurate test signals for testing and calibrating BTSC system receivers and transmitters. The complex, multiband noise reduction system used in the BTSC system is simulated mathematically. Then test signals are digitally synthesized, providing precise amplitude and phase relationships. The RE540 is fully programmable over the IEEE-488 bus. **RE Instruments Corporation, Westlake, Ohio. INFO/CARD #111.**

Miniature Circulator High Power

A new line of miniature microwave isolators and circulators permits transmitted power handling capacity up to one kilowatt (1 kW) peak but retains the small size previously limited to lower power ratings. Individual units having a center frequency of between 380 MHz and 6.0 GHz will operate over a wide temperature range of −40 to +70 deg. C without degradation of performance. Typical electrical characteristics include 35 dB or greater isolation and 0.7 dB insertion loss even at 50 watts mean power. **Nore Microwave Limited, Southend-on-Sea, Essex, U.K. INFO/CARD #110.**

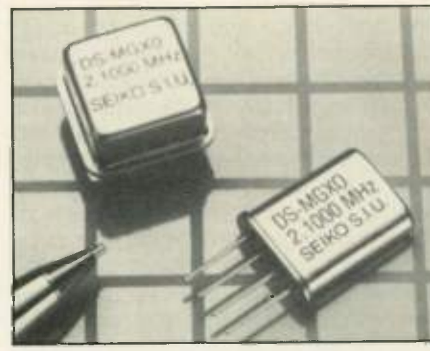
RF Coils and Chokes are Conformal Coated

West-Cap Arizona has introduced a full line of conformally coated axial RF coils and chokes. These new products, the series 1AC and 20000C, were previously available only in molded form. Automatic insertion of these new inductor products is available with axial tape and reeling per

EIA specification RS296D. Available inductance values range from 0.1 to 1,000 microhenries. **West-Cap Arizona, subsidiary of SFE Technologies, Tucson, Ariz. INFO/CARD #109.**

High-Precision Miniature Oscillators

Seiko Instruments has introduced new crystal oscillators, the MGXO-DM and MGXO-SM Series, that combine high precision and low power consumption with greatly reduced size. The 2.1 MHz models incorporate a CT cut crystal unit, an electric charge injection variable capacitance device and a CMOS IC. These oscillators are capable of changing capacitance continuously and electrically, rather than mechanically, and can hold a capacitance value for an extended period of time. As a result, they permit



high-precision external frequency adjustment by voltage control. THe MGXO crystals are $78.00 each (50s). **Seiko Instruments, U.S.A, Inc., Torrance, Calif. INFO/CARD #217.**

Counter Timers Measure Up to 1 GHz

O.K. Industries has added two new Universal Counter/Timers for 100 MHz to 1 GHz to its line of test instrumentation. Designated Models 5010 (100 MHz) and 5110 (1 GHz), the counter/timers incorporate frequency, period, period average, time interval, time interval average, frequency ratio and totalize measurement modes. There is an eight-digit 0.56" red LED display with six additional annunciators for external standard, gate, overflow, MHz/us, kHz/ms and seconds. Both models have a high stability time base option based upon a fast warm-up ovened crystal. List prices are $1,195 for Model 5010 (100 MHz) and $1,395 for Model 5110 (1 GHz). **O.K. Electronics Division, Yonkers, N.Y. INFO/CARD #216.**

Thick Film Switch Operates to 1 GHz

This new unit features low video feedthrough and small size. The hermetically sealed unit is 1" × 1" × .25" and operates at 1.1 GHz. Switching spikes are suppressed to less than 1 volt peak, 2.5 nanosecond maximum width. Other model RF switches cover from .5 MHz to 6.0 GHz. **Tektron Micro Electronics, Inc., Hanover, Md. INFO/CARD #215.**

Kit Contains 20 Coax Adapter Cables

TPI-5000 Kit includes 20 cables, each a different combination of BNC, TNC, N, UHF and Mini-UHF connectors. Special RG-58 A/U cables are very soft and easy to handle, and are colored bright yellow. Connectors are machined brass with silver plated contacts. The kit is furnished with two wall racks, and priced $125 complete. **Test Probes, Inc., La Jolla, Calif. INFO/CARD #108.**

Digital Signal Processing (DSP) Course is PC-Based

Rapid Systems announces the R100 series of digital signal processing courses. Priced at $1,299 to $3,299, depending on included hardware, the course includes a custom 250-page DSP text coupled with the hardware/software to turn a PC into a digital oscilloscope, FFT spectrum analyzer, data logger, or data acquisition module. Courses are available for Apple and C-64 computers, with chapters covering data acquisition, sampling, digital filtering, FFT and spectrum analyzers, sample and holds, aliasing and windowing, and A to D converters. **Rapid Systems Inc., Seattle, Wash. INFO/CARD #151.**

Program Optimizes Impedance Matching Networks

OptiMatch is a full-blown impedance match network optimization program for Apple and IBM computers. With it, the user can optimize either single ended or complex interstage networks for lowest possible VSWR across a band, or to create models of devices using measured S-parameter data or impedance measurements. The user can also optimize the noise match for an LNA. It has a full feature file editor and will allow the user to change the optimization algorithm itself to "fine tune" it for special designs. The price is $149.50 for the Apple and $189.50 for the IBM fast machine code version. **Microwave Software, San Juan Capistrano, Calif. INFO/CARD #150.**

New Version of SUPER-COMPACT Enhances Speed

Compact Software announces an enhanced version of SUPER-COMPACT to be released in Revision 1.9 using a dedicated sparse matrix technique. Standard practice has been to use the Gauss-Jordan matrix inversion in which the execution speed is essentially proportional to one over the square of the number of nodes. While SUPER-COMPACT allows circuit execution either in two-port or nodal analysis, the nodal analysis is a must for three- and four-port circuits as found in microwave circuits like couplers. Circuits above 10 or 15 elements described in such a way show slow execution time. By using a nodal sparse matrix technique, the execution time is essentially linearly proportional to the number of components. This method then becomes equally fast as the two-port analysis which uses a special form of sparse matrix. **Compact Software, Inc., Paterson, N.J. INFO/CARD #149.**

ACAP Analyzes AC Circuits

ACAP, from RF Engineering, is a "stripped down" version of LCAP. It's like LCAP but without time domain analysis, and library active components. It retains features that make it a powerful circuit analysis program. For example, ACAP can analyze a twelve node RF circuit for attenuation (S21), return loss, phase, group delay, and impedance, at a rate of 1 frequency per second, using a standard IBM-PC without 8087 coprocessor. Features include full screen editing, matrix analysis for unlimited circuit topologies, transistor models accurate to 1000 MHz, inductors and capacitors of any Q. Circuits can be saved on disk. ACAP is priced at $49.00. ACAP-87 for the 8087 coprocessor is $90.00. **RF Engineering, Norwich, N.Y. INFO/CARD #148.**

Program Designs and Analyzes Combline Filters

EEsof, Inc.'s latest program for IBM PC and PC-compatible computers allows even a novice microwave engineer to design precise combline filters. Filter III: Combline Design™ (CLD) is the third in a series of PC-based microwave filter design programs by veteran filter design consultants Bob Wenzel and Bill Erlinger.

The program enables the design and realization of combline bandpass filters with exact equal-ripple passband response, using constant diameter rods. CLD designs broad bandwidth (2:1) filters with 2 to 20 resonators, configured with the capacitive loading in either the cover or the resonator. It creates equivalent circuits containing all line element impedances as well as resonator loading capacitor values. The program creates design files, S-parameter files, and analyzed response tables. Design files include output for equal diameter rod dimensions, requires capacitive loading elements, tap position, tap line dimensions, and line element impedances. CLD sells for $7,500; quantity discounts are available. **EEsof, Inc., Westlake Village, Calif. INFO/CARD #147.**

Software Package Features Active Filter Design

RLM Research announces a major revision to the "Active Filter Design" software package for the IBM family of personal computers. This new release (version 2.00) adds an interactive graphics facility that allows plotting the gain, phase or group delay as well as the impulse or step response. The response from individual sections or different filters may be superimposed on the screen. Four on-screen cursors allow real time measurement of the value of any point on the screen or defining an area to be expanded to the full screen size. Finished plots can be printed on a dot matrix printer or plotted on any HP compatible plotter. The "Active Filter Design" program designs Butterworth, elliptic, Chebyshev and Bessel low pass, high pass, band pass and band stop active filters. It also allows the direct entry of pole and zero locations or transfer functions and can convert low pass prototype poles and zeros to the desired filter configuration. This revision is being sent to all owners of version 1.01 at no charge. The price of the program remains at $450. **RLM Research, Boulder, Colo. Please circle INFO/CARD #146.**

Engineering Software Catalog

Interleave, Inc. announces a new mail order engineering software catalog. The catalog is directed towards the electronic design engineer, but also holds a wide variety of engineering related software covering eight specific application areas — CAD, electronics, graphics, mathematics, engineering support, project management, scientific (data acquisition), statistics and text processing. The catalog is designed to offer a convenient source of software programs for managing and design engineers. Each product is fully described in detail. An engineer or scientist would have to read the ads in more than fifteen periodicals every month in order to accumulate a comparable amount of information. This 32-page catalog has comprehensive descriptions of over 70 important products in one place. **Interleave, Inc., Cleveland Heights, Ohio. Please circle INFO/CARD #145.**

rf literature

Catalog Features HF-to-Microwave Feedlines, Antennas

The new Andrew catalog features a systems planning section, information on new product lines and coverage of all the additions to Andrew telecommunications products since the last catalog was issued, including the new Andrew HF antenna products acquired from Granger. Catalog No. 33 also has information on terrestrial and earth station antennas, broadcast antennas, radar and navigational aid antennas, HELIAX™ cable waveguides, RADIAX™ radiating coaxial cables, optical fiber systems, plus GRASIS™ towers and equipment shelters. **Andrew Corp., Orland Park, Ill. INFO/CARD #142.**

Microwave Amplifier Brochure

A four-page color brochure detailing its microwave amplifier capabilities is available from M/A-Com Omni Spectra, Inc. In addition to outlining typical frequency range, power output, noise figure data and other capabilities, the brochure introduces M/A-Com's NO-PAK™ technology, where custom and standard GaAs FET amplifiers are provided in packageless form for incorporation directly into microwave integrated circuit assemblies. **M/A-Com Omni Spectra, Inc., Tempe, Ariz. INFO/CARD #141.**

RF, IF and Mixer Amplifier Catalog

TIW Systems announces the availability of a 12-page catalog describing its line of RF, IF and mixer amplifiers, designed to meet high performance specifications. The catalog features TIW Systems' standard fixed and variable gain RF amplifiers and mixer amplifiers, plus new products including fixed gain high power and voltage-controlled variable gain RF amplifiers, IF preamplifiers, IF buffer amplifiers, IF limiting amplifiers, and DC through 400 MHz video amplifiers. **TIW Systems, Inc., Sunnyvale, Calif. INFO/CARD #140.**

Full-Line Catalog from Narda

Catalog 24, Narda's new full-line catalog, contains descriptions, specifications and outline drawings for more than 700 microwave component and instrument products, including attenuators, couplers, power dividers, PIN switches, sources, MICs, radiation and power monitors, microwave measurement and path alignment systems. Narda's state-of-the-art operation is documented in a new capabilities section. **The Narda Microwave Corporation, Hauppauge, N.Y. INFO/CARD #139.**

Brochure Notes Hi-Rel Capabilities

Merrimac Industries presents an applications book on its Hi-Rel products, programs, and capabilities. Products described include Hi-Rel signal processing components produced over the past 20 years, plus the latest supercomponents being used in current Hi-Rel and Mil-Screen programs. **Merrimac Industries, Inc., West Caldwell, N.J. INFO/CARD #138.**

Brochure Describes RF and EMC Services

A new brochure from Comsearch Applied Technology (CAT) describes the capabilities of the company. CAT divisions address problems related to radio frequencies (RF) and electromagnetic compatibility (EMC). Services include TEMPEST applications, secure environments, EMC design and testing, research and development and training. Designers have the know-how to solve RF problems at the board level or at the system level. Technical competence ranges from field installation to testing for human exposure to radio frequency energy hazards. **Comsearch, Inc., Reston, Va. INFO/CARD #137.**

Coaxial Products Catalog

Catalog #986 features a full line of coaxial adapters, connectors, attenuators, terminations, and cable assemblies. Cable assemblies featured are both flexible and semi-rigid. Also featured are twinax connectors, twinax adapters, and hex crimping tools. Pricing on over 1,000 standard catalog items as well as technical specification are included. Special length assemblies as well as special coaxial cable assembly configurations are available upon request. **Pasternak Enterprises, Irvine, Calif. INFO/CARD #136.**

Technical Data Describes Parallel Plate Chip Caps

Tecdia, Inc. has available catalog, technical data and applications information on their line of single layer parallel plate chip capacitors. Available in single value and binary configurations, Tecdia capacitors are designed for thin and thick film MIC applications up to 40 GHz. The technical manual and applications guide includes complete data for including these products in MIC or hybrid designs. **Tecdia, Inc., Mountain View, Calif. INFO/CARD #135.**

Mixer and Frequency Translation Components

A 28-page brochure describes complete double balanced mixer specifications for .1-26 GHz applications. Units covered include mixer modules, connectorized mixers, quadrature IF mixers, SAT-COM TVRO mixers, miniature wideband and biasable double-balanced mixers, plus microwave IF and waveguide balanced mixers. **Norsal Industries, Inc., Central Islip, N.Y. INFO/CARD #134.**

Fault Location System Data Sheet

A new data sheet describes Wiltron's second generation 5600 series option P2-FF distance-to-fault location system. The second generation provides the same superior performance of the first transmission line test and fault location system but in a smaller package and in a broader frequency range. The system's new smaller, lighter enclosure offers a 50 lb. weight reduction over the previous model. With a new wider frequency range the system can test up to 40 GHz. **Wiltron Company, Morgan Hill, Calif. INFO/CARD #133.**

Microwave Substrate Selector Chart

A brochure featuring Di-Clad® low loss microwave substrate materials for stripline and microwave circuitry application is available from Keene Laminates. The easy-to-read selector chart contains information on each grade of Di-clad, including dielectric constant, dissipation factor, standard dimensions and physical descriptions of each substrate. A graph illustrates the controlled dielectric constant for each Di-Clad grade. Additional information regarding substrate thickness, copper cladding and dielectric constant is also presented in quick-reference chart form. **Keene Laminates, East Providence, R.I. INFO/CARD #132.**

Inductor Guide and Cross Reference

"Fixed Inductor Selector Guide and Cross Reference" has been published by J.W. Miller Division of Bell Industries. Over 1100 inductance values with current and resistance ratings are listed from .01µH to 500,000µH. The Cross Reference section contains the part numbers of seven manufacturers and Mil Spec listings, with the corresponding J.W. Miller part numbers. **J.W. Miller Division, Bell Industries, Rancho Dominguez, Calif. INFO/CARD #131.**

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Capacitance Cj (pF) MAX	.03	.06	.02	.07
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Switching Time TS (ns) TYP	5	5	15	15

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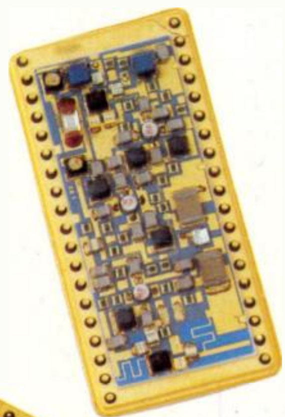
INFO/CARD 100

SAW Questions and Andersens.

Oscillators:

QUESTION: I've heard a lot about SAW hybrid oscillators. When should I consider using one?

ANDERSEN: Any time you need a compact, rugged low noise source in the 100 MHz to 1 GHz range.



QUESTION: What do you mean by low noise?

ANDERSEN: We build SAW hybrid oscillators to phase noise specs of −95dBc/Hz at 100 Hz offset from the carrier to less than −160dBc/Hz at 100KHz.

QUESTION: Are there high NRE costs associated with SAW oscillators?

ANDERSEN: Not with us. We've developed designs for fixed frequency and VCO applications which can cover the entire 100 MHz to 1 GHz range. So we normally don't charge any NRE. That's why we're the leaders in SAW oscillators.

QUESTION: Where are SAW oscillators being used?

ANDERSEN: Our fixed-frequency SAW oscillators are being used in IFF and air traffic control systems, as local oscillators for frequency up/down converters in radar and communications systems, as master clocks for digital communications systems, just to name a few of the applications.

Our SAW VCO's are used in a wide range of Phase Locked Loops including timing recovery circuits for fiber optic data transmission systems where the oscillator reestablishes the timing of the data stream at the end of the fiber optic lines.

QUESTION: How do I get started?

ANDERSEN: Just give us a call. We'll do everything we can to meet your needs. If you simply want more information, send for our handbook on SAW Hybrid Oscillators. Write to Andersen Laboratories, 1280 Blue Hills Avenue, Bloomfield, CT 06002, Phone (203) 242-0761/ TWX 710-4253 2390

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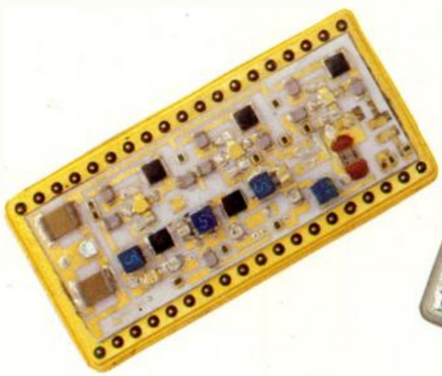

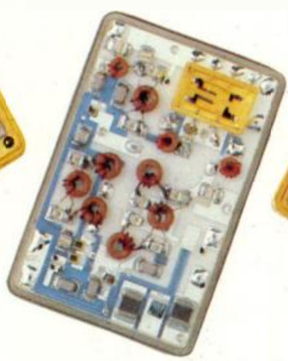

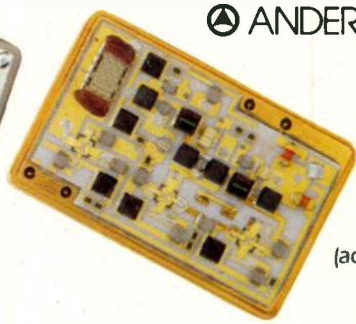

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