July/August 1983

A Cardiff Publication

r.f. design

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900 MHz	MOBILE APPLICATIONS				
Type	Power (W)	Freq. (MHz)	Gain (db) min.	VCC (V)	Package
BLU98	0.5	900	9.0	12.5	SOT-103/E1
BLV90	1	900	7.5	12.5	SOT-172
BLV91	2	900	6.5	12.5	SOT-172
BLU99	4	900	7.3	12.5	SOT-122
BLV92	4	900	8.0	12.5	SOT-171
BLV93	8	900	6.0	12.5	SOT-171
BLV94	15	900	6.0	12.5	SOT-171

400 to 512 MHz	MOBILE APPLICATIONS				
BLU60/12	60	175	4.8	12.5	SOT-119
BLU45/12	45	175	5.1	12.5	SOT-119
BLU30/12	30	175	6.0	12.5	SOT-119
BLU20/12	20	175	6.5	12.5	SOT-119
BLW82	30	175	5.0	12.5	SOT-119
BLW81	10	175	6.0	12.5	SOT-122
BLU99	5	175	10.5	12.5	SOT-122
BLW80	4	175	8.0	12.5	SOT-122
BLW79	2	175	9.0	12.5	SOT-122
BLX65	2	175	6.0	12.5	TO-39

175 MHz	MOBILE APPLICATIONS				
BLV75/12	75	175	7.0	12.5	SOT-119
BLV45/12	45	175	6.5	12.5	SOT-119
BLV30/12	30	175	8.2	12.5	SOT-119
BLW60C	45	175	5.0	12.5	SOT-120
BLW31	28	175	9.5	12.5	SOT-120
BLY89C	25	175	6.0	12.5	SOT-120
BFQ43	4	175	12.0	12.5	TO-39E
BFQ42	2	175	10.5	12.5	TO-39

30 to 900 MHz	BASE STATIONS				
Type	Power (W)	Freq. (MHz)	Gain (db) min.	VCC (V)	Package
BLW96	200	30	13.5	50	SOT-121
BLV25	175	108	10.5	28	SOT-119
BLV80/28	80	175	6.5	28	SOT-119
BLV33F	85	225	10.5	28	SOT-119
BLV36	120	225	10.0	28	SOT-161
BLU53	100	400	6.5	28	SOT-161
BLV97	30	860	6.5	24	SOT-171
BLV57	38	860	6.5	25	SOT-161

66 to 870 MHz	AMPLIFIER MODULES FOR LAND MOBILE				
Type	Freq (MHz)	P In (MW)	P Out (W)	VCC	Package
BGY32	68-88	100	20	12.5	SOT-132
BGY33	80-108	100	20	12.5	SOT-132
BGY35	132-156	150	20	12.5	SOT-132
BGY36	148-174	150	20	12.5	SOT-132
BGY43	148-174	150	13	12.5	SOT-132B
BGY40A	400-440	100	7.5	12.5	SOT-132C
BGY41A	400-440	150	13	12.5	SOT-132C
BGY40B	440-470	100	7.5	12.5	SOT-132C
BGY41B	440-470	150	13	12.5	SOT-132C
BGY40A	470-512	100	7.5	12.5	SOT-132C
BGY41C	470-512	150	13	12.5	SOT-132C
BGY45A	68-88	150	30	12.5	SOT-301-A-03
BGY45B	144-175	150	30	12.5	SOT-301-A-03
BGY46A	400-440	30	1.5	9.6	SOT-26NC
BGY47A	400-440	45	2.2	9.6	SOT-26NC
BGY47B	430-470	45	2.2	9.6	SOT-26NC
BGY47C	460-512	45	2.2	9.6	SOT-26NC
BGY22	380-512	50	29	12.5	SOT-75A
BGY23	380-480	2.5 WATTS	7	12.5	SOT-75A

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INFO/CARD 1


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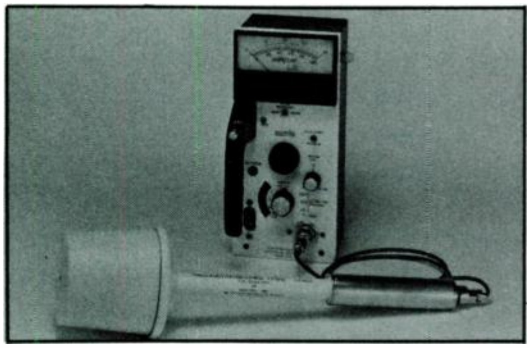
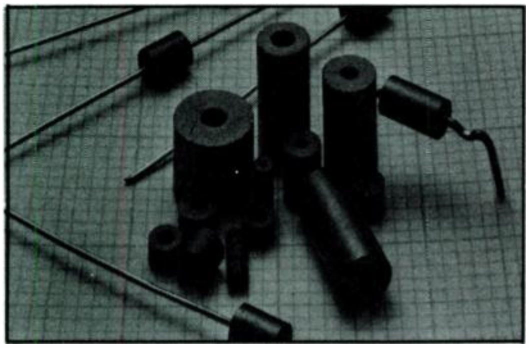
rf design

July/August 1983

Cover Story, P. 12



RF Radiation Hazards, P. 24



Ferrite Beads, P. 39

Cover

Features

Departments

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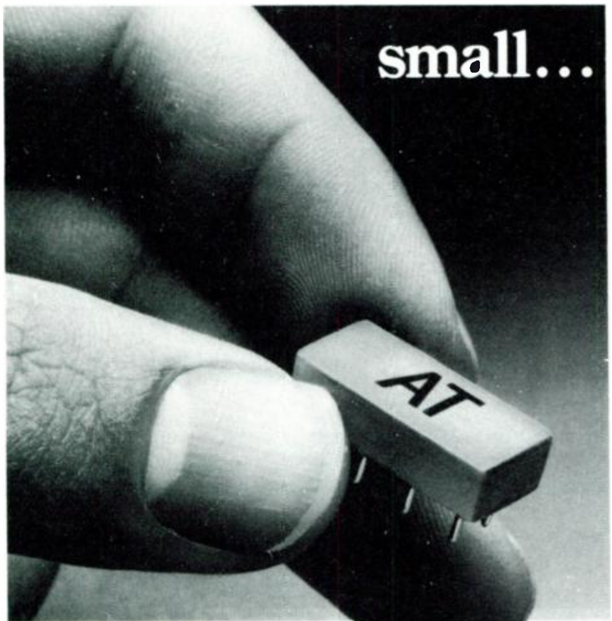
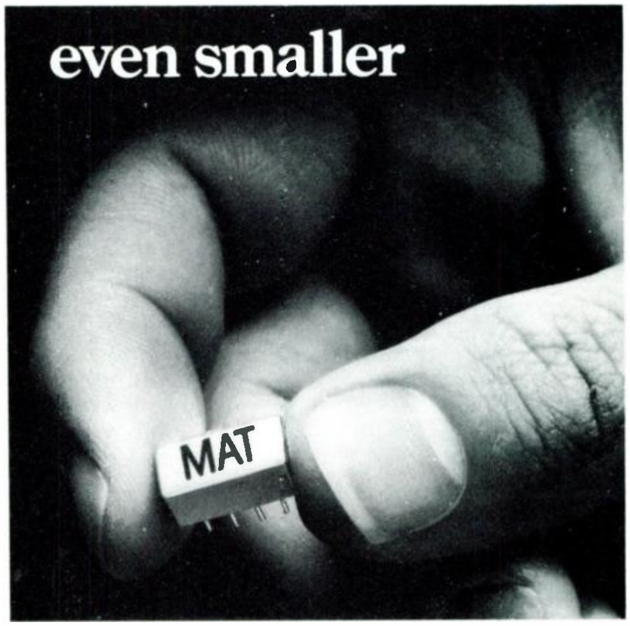
plug-in attenuators





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AT-SERIES	MAT-SERIES	Attenuation, dB Nominal	Attenuation, dB Tolerance	Attenuation Variation With Frequency DC-1000	Attenuation Variation With Frequency 1000-1500	VSWR Max DC-1000	VSWR Max 1000-1500
AT-3	MAT-3	3	±0.2dB	0.6dB	1.0dB	1.3:1	1.5:1
AT-6	MAT-6	6	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
—	MAT-9	9	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
AT-10	MAT-10	10	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
—	MAT-12	12	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
—	MAT-15	15	±0.3dB	0.6dB	0.8dB	1.3.1	1.5:1
AT-20	MAT-20	20	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1
AT-40*		40	±0.3dB	0.6dB	0.8dB	1.3:1	1.5:1

*AT-40 Frequency Range DC-500 MHz
Frequency Range: DC-1500 MHz, power (max.) MAT = 0.5W, AT = 1W

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INFO/CARD 5

Ruth Breashears
One That Made Things Happen

By Bill W. Childs
Publisher

In our lives we occasionally meet, work with, and even get to know some really wonderful people. Sometimes we do not fully appreciate the experience until it is too late. How many times have you heard someone say, "I wish I had gotten to known someone better while I had the opportunity"?

Ruth Breashears, our *r.f. design* sales manager, has accepted a position in California. Usually, I am one who wishes people the very best when they go on to new challenges. However, I must admit I did ask Ruth to reconsider leaving Denver and *r.f. design* magazine for Sunnyvale, California. In the end, though, the San Fransico Bay area prevailed.

I first met Ruth two years ago when she came to Cardiff as sales manager for *r.f. design*. She almost single handedly tried to pick up the pieces of a great idea, on a poorly managed magazine. Her tenacity and long hours were the talk of the office. In February of last year, I joined her in the effort and we shortly brought on Larry Brewster as editor, Tami Baird as editorial assistant, Sue Myers and Blair Sutfin as advertising space salesmen.

Through all of this, Ruth's tenanciousness endured. In addition to advertising sales, Ruth's involvement encompassed the entire magazine. She tracked down cover stories, bird dogged late authors and worked with the artists to bring the reader better graphic presentations.

Ruth chose the cover for the July/August issue of *r.f. design* and we dedicate this issue to her. I know she has many friends in this business that know and love her as our staff does. I am sure those of us who know her all wish her the very best in her new endeavors.

For those of you who have never had the privilege of meeting Ruth, I would say that if you have been a loyal reader over the past two years, you have met her in her own special way. I know she has had a wonderful time bringing *r.f. design* to its readers and advertisers and she would love to have met you all. For two years we have not been able to separate Ruth from r.f. or r.f. from Ruth; they were one and the same.

She leaves, knowing she has left a seasoned and enthusiastic staff. We all look forward to continuing our efforts to produce a good magazine for you, the *r.f. design* reader. I thought you might like to meet a lady who played a major role in making it possible.

Sue Myers, who has been in advertising sales for *r.f. design* under Ruth, will assume the position of sales manager. Also this summer Ken Kadash will join the *r.f. design* staff as Associate Publisher. Ken will step into the role of sales and editorial management as the year progresses. In fact later on this summer both Larry Brewster and Ken will be traveling around the country in search of good editorial material for 1984.

If you know of some outstanding material or potential articles please give Ken or Larry a call. They would be very pleased to discuss the editorial content of the magazine for the future.

Please stop by and see us at our booth at the Wescon show November 8-11.

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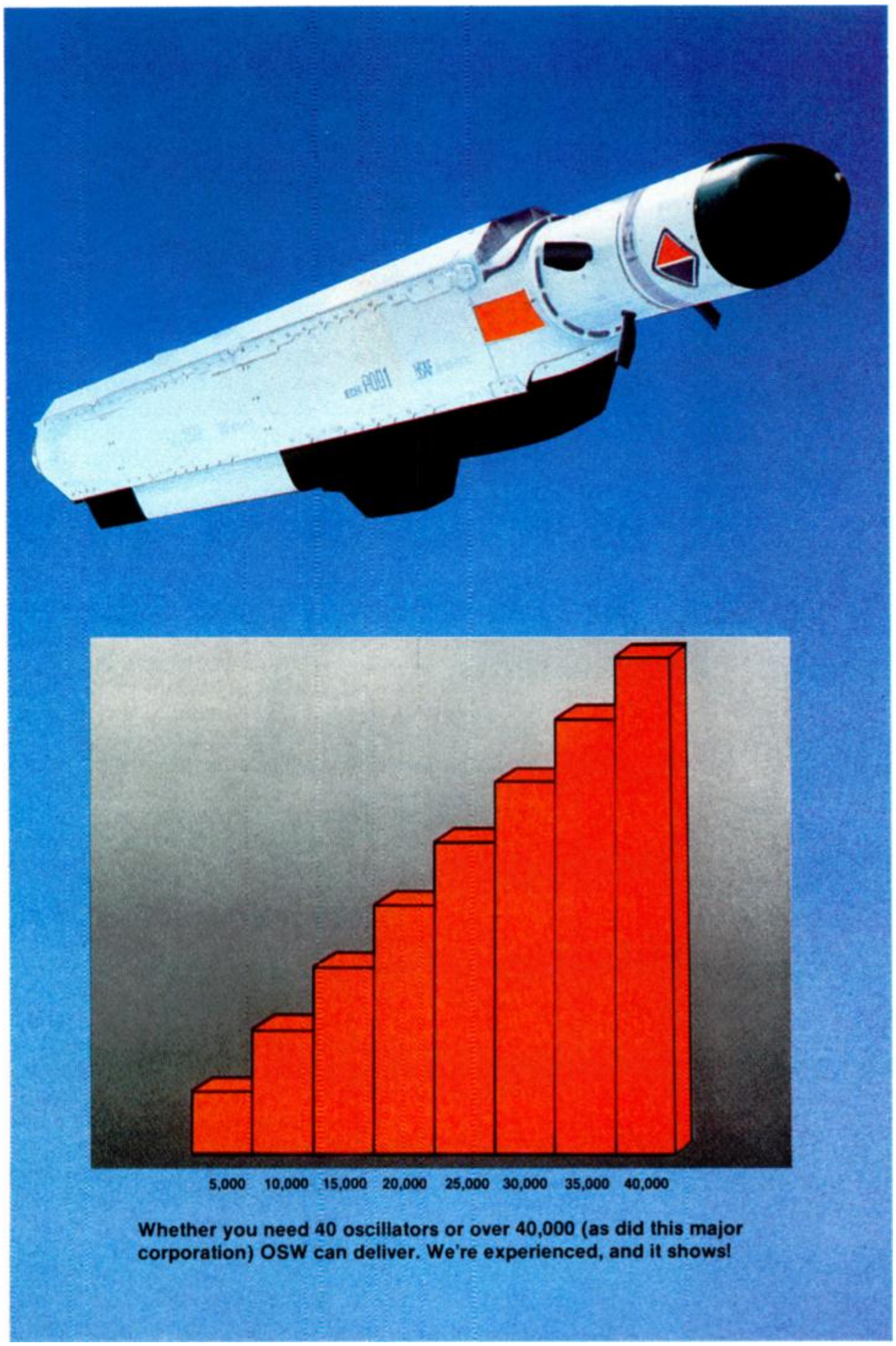

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INFO/CARD 6

"Calculator Aided Stub Matching" Article Errors

Dear Larry:

My compliments on a job well done in presenting my manuscript "Calculator Aided Stub Matching" (May/June 1983). One confusing error did manage to sneak through and I would like to inform your readers of the following:

Page 31, 1st and 2nd paragraphs: Please replace $B(x)/Y_o$ with $G(x)/Y_o$.

The remaining part of the 2nd paragraph should read: Using equation (1a), equation (3), and equation (5), with G_T/Y_o replaced by G_T/Y_o new.

Also, page 42 contains a printout for both programs USTRIP and MAT and not just USTRIP as shown in the title heading.

Keep up the excellent work.

Sincerely,
Alan Victor

Dear Mr. Brewster:

Thank you for the many fine articles which have appeared in your magazine. I am glad that you have taken up the slack, and filled the void created when *MicroWaves* ceased to publish calculator based design articles.

Unfortunately, to err is human, and by this I refer to the article entitled "Calculator Aided Stub Matching" which appeared in the May/June issue. The article contains several errors which are readily apparent when one compares the calculator program with the design equations. In addition, and most important of all, the expression given for the quantity B(x)/Yo in equation (7) is incorrect; due to the error in this equation, the stub lengths calculated in (8) are off by a small amount in certain matching instances. The error, thank goodness, is small, but it should not be there in any event. One may easily verify the errors present in the design example on pages 31 and 33 by either hand calculation, or by the program COMPACT which is available on nationwide time-share services. When a correct equation is used for the quantity B(x)/Ro, the calculator program generates the attached table of values for the in/out match networks in the example problem.

Please continue to publish this type of article. Although errors can happen, especially if one is a bit hasty to submit his work without thoroughly reviewing it, the work done by Mr. Victor is invaluable and has contributed to the advancement of the design art.

Sincerely,
James J. Lev
Manager, RF/Microwave Design
Cartwright Engineering, Inc.

INPUT

Zo(Ω)	series line λ	OSTB λ	SSTB λ
50	.122	.180	---
	.002	---	.070
75	.082	.202	---
	.001	---	.048
100	.062	.213	---
	.001	---	.037

OUTPUT

Zo(Ω)	series line λ	OSTB λ	SSTB λ
50	.228	.224	---
	.178	---	.026
75	.205	.225	---
	.158	---	.025
100	.185	.226	---
	.141	---	.024

NOTE: The two tables of tabulated data shown above are the correct solutions to the 2N3570-750MHz design example given in the text on pages 31 and 33.

Comments on "Triple Tuned Circuits" Article

Dear Sir:

The March/April issue of *r.f. design* contains a letter from Mr. Tony Tortorella commenting upon an article by Mr. Andrzej Przedpelski titled "Triple Tuned Circuits". I do not have Mr. Przedpelski's article at the present time, and will address my comments to the problem formulated by Mr. Tortorella. This is the design of a third-order Butterworth bandpass filter with passband extending from 20 KHz to 60 KHz, to be driven from a 25 ohm source and operating into a 75 ohm load. The solution proposed by Mr. Tortorella is to make use of a partitioning theorem related to Bartlett's Bisection Theorem. This method yields a network which exhibits the ideal Butterworth bandpass characteristic when terminated as specified, but the network is not matched to source and load. This partitioning and scaling method applied is particularly useful for lowpass filters which must operate between unequal terminations, since a lowpass filter, by virtue of the fact that it must be essentially transparent down to DC, cannot contain an impedance matching transformer. A bandpass filter, on the other hand, can usually contain an ideal transformer, and this can frequently be accomplished through the addition of a single capacitor.

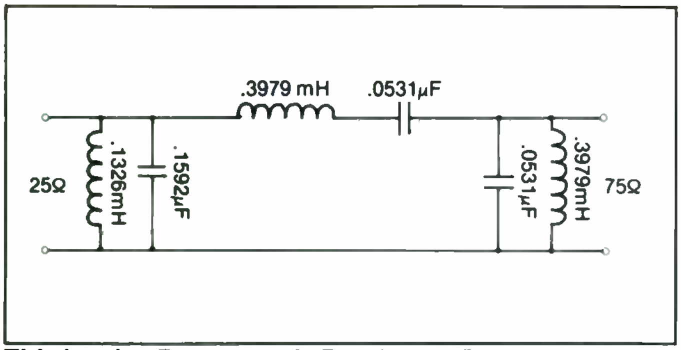
To illustrate this point, I enclose schematic diagrams of two third-order Butterworth bandpass filters. The first is the circuit derived from partitioning, which is discussed by Mr. Tortorella. The second is the solution incorporating an ideal transformer, which requires an additional capacitor. Also enclosed are frequency analyses of these two networks derived from a computer program which analyzes ladder networks. For purposes of this illustration, the reactive elements have been assumed lossless. The column headed "ATTEN." is the actual transducer loss of the network. Three points become evident in comparing the two realizations. First, both exhibit the ideal expected Butterworth characteristic in amplitude and phase, but the partitioned network includes a flat loss of about 1.25 dB which must be subtracted from all loss readings to show relative attenuation. The frequency of transparency is the geometric mean frequency of 34.64 KHz (not 36.64 KHz). Both networks show essentially zero insertion phase change at this frequency, and the input impedance has essentially zero angle (is resistive). The partitioned network, however, shows an impedance of 75 ohms, being purely transparent to the load, and a return loss of only about 3 dB. The transformed network shows an input impedance of 25 ohms and a return loss of nearly 84 dB because of the ideal transformer which it contains.

The intent of this correspondence is to point out to any readers not highly versed in filter design methods that there is a difference between a network designed to operate between unequal impedances without matching and one which is designed to match a source to an unequal load. Furthermore, the single extra capacitor required for the matched case will, in many applications, "pay its own freight" in terms of improved power transfer, improved return loss, and lower sensitivity to element value variations.

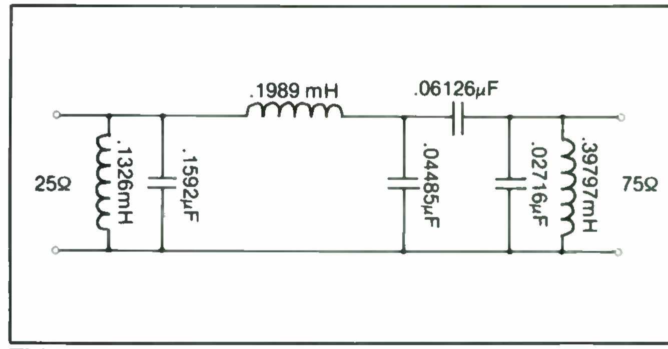
Sincerely,
David C. Bidwell, Sr. Eng.
RFL Industries Inc.,
Boonton, NJ 07005

Reader Bidwell's letter contains useful information. It also shows that the articles and letters are well read. I think you should encourage this type of interchange of ideas: it gives the readers the feeling that it is "their" magazine, since they can communicate their ideas; and also it can provide high level technical information at a more informal level. Note the difference in your letters and the "political" letters of some of the other magazines.

Best regards,
Andrzej B. Przedpelski
Vice President, Development
A.R.F. Products, Inc.



Third order Butterworth Bandpass Filter Partitioned (mismatched).



Third order Butterworth Bandpass Filter Transformed in Impedance (matched).

FREQ.	ATTEN.	PHASE	Z IN (MAG)	Z IN (ANG)	Y IN (MAG)	RETURNLOSS
14640.00	15.0014	163.2856	16.639954	89.002467	60.09632E-03	13.95095E-02
17140.00	9.2631	137.6494	23.081411	86.377752	43.32491E-03	54.77377E-02
19640.00	4.7504	219.8934	32.796574	77.958894	30.49099E-03	17.71381E-01
22140.00	2.2706	255.3496	41.603481	61.500944	24.03645E-03	39.02429E-01
24640.00	1.4507	285.6781	44.764406	49.081325	22.33918E-03	54.67164E-01
27140.00	1.2759	309.3290	49.946764	42.077085	20.02132E-03	59.41966E-01
29640.00	1.2513	328.5246	59.358826	33.064966	16.84669E-03	60.14741E-01
32140.00	1.2495	345.0769	70.034192	18.657579	14.27874E-03	60.20141E-01
34640.00	1.2494	0.0003	75.001875	-0.006068	13.33300E-03	60.20437E-01
37140.00	1.2494	13.8763	70.627276	-17.499870	14.15884E-03	60.20592E-01
39640.00	1.2500	27.1022	62.137560	-29.986433	16.09333E-03	60.18656E-01
42140.00	1.2562	40.0067	54.527197	-37.823923	18.33947E-03	60.00336E-01
44640.00	1.2824	52.8845	49.188098	-42.795277	20.33012E-03	59.23162E-01
47140.00	1.3581	65.9811	45.965833	-46.614246	21.75529E-03	57.10008E-01
49640.00	1.5292	79.4409	44.226613	-50.734989	22.61082E-03	52.75387E-01
52140.00	1.8526	93.2329	43.001851	-56.115077	23.25481E-03	45.93533E-01
54640.00	2.3797	107.1018	41.216518	-62.710698	24.26212E-03	37.48219E-01
57140.00	3.1356	120.6082	38.379619	-69.467331	26.05550E-03	28.88529E-01
59640.00	4.1052	133.2745	34.877277	-75.248300	28.67196E-03	21.36591E-01
62140.00	5.2409	144.7495	31.364947	-79.604694	31.88273E-03	15.43857E-01
64640.00	6.4820	154.8840	28.231844	-82.670623	35.42099E-03	11.05885E-01

ENTER 1 FOR NEW FREQ., 2 FOR NEW DATA, 3 TO END? 3

Network Derived by Partitioning

FREQ.	ATTEN.	PHASE	Z IN (MAG)	Z IN (ANG)	Y IN (MAG)	RETURNLOSS
14640.00	13.7519	163.2850	18.594826	88.711917	53.77840E-03	18.70307E-02
17140.00	8.0136	187.6484	28.563250	85.035627	35.01002E-03	74.68563E-02
19640.00	3.5006	219.8924	51.130231	68.635330	19.55790E-03	25.69755E-01
22140.00	1.0207	255.3497	63.573625	16.341089	15.72979E-03	67.89114E-01
24640.00	0.2009	285.6794	37.817662	-6.850489	26.44267E-03	13.44763E+00
27140.00	0.0262	309.3311	28.186009	-5.641925	35.47859E-03	22.20667E+00
29640.00	0.0017	328.5270	25.529164	-1.949153	39.17089E-03	33.98969E+00
32140.00	0.0000	345.0794	25.030551	-0.248574	39.95118E-03	52.94274E+00
34640.00	0.0000	0.0028	25.000658	-0.007250	39.99895E-03	83.79215E+00
37140.00	0.0000	13.8786	25.020311	0.189421	39.96753E-03	55.38006E+00
39640.00	0.0007	27.1044	25.292435	1.305936	39.53751E-03	37.85949E+00
42140.00	0.0069	40.0087	26.310292	3.495644	38.00794E-03	28.00446E+00
44640.00	0.0331	52.8862	28.721921	6.052911	34.81661E-03	21.19466E+00
47140.00	0.1088	65.9826	33.348476	7.485744	29.98638E-03	16.06439E+00
49640.00	0.2800	79.4421	41.232723	5.584663	24.25258E-03	12.04557E+00
52140.00	0.6034	93.2339	53.053366	-2.665279	18.84895E-03	88.70281E-01
54640.00	1.1305	107.1025	65.330463	-20.059413	15.30679E-03	63.97996E-01
57140.00	1.8864	120.6086	67.001064	-43.017234	14.92514E-03	45.30670E-01
59640.00	2.8560	133.2748	57.516600	-61.435884	17.38628E-03	31.70301E-01
62140.00	3.9916	144.7497	47.005924	-72.515162	21.27391E-03	22.10368E-01
64640.00	5.2327	154.8842	39.016106	-78.811570	25.63044E-03	15.47347E-01

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FREQUENCY RANGE (MHz)	GAIN (dB)	NOISE FIGURE (dB)	THIRD ORDER INTERCEPT (dBm)	POWER OUT (1 dB COMPRESSION) (dBm)	MODEL NO.
10-200	8	1.3	+30	+11	AM-117
5-500	10	3.5	+42	+22	AM-123
10-1000	11	4.5	+35	+17	AM-145
200-1000	12	2.2	+20	+ 6	AM-142
300-1800	12	1.9	+19	+ 7.5	AM-153
150-5200	15	6.0	+30	+19	AM-250

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INFO/CARD 8

RF Radiation Hazards Measurement and Evaluation

Equipment designers and users should be aware of potential risks, pertinent standards, and measurement devices associated with RF radiation.

By Alan Schwartz
The Narda Microwave Corp.
Hauppauge, NY 11788

The number of applications for RF technology continues to grow at an ever increasing rate. In addition to the expansion of communications capabilities to keep pace with our insatiable needs for bandwidth and the advancement of the state of the art in radar and electronic warfare systems, industrial, scientific and medical applications for RF are on the increase as well. RF energy is used in industrial processes such as plasma etching of semiconductors, welding and brazing of metals, laminating glueing and drying of wood and other products, as well as heating, sealing and trimming of plastics.

Scientific applications include "pumping" of particle accelerators, and the medical sector uses RF for therapeutic purposes, such as diathermy and for RF hyperthermia as a treatment for cancer. This is by no means an exhaustive list of modern applications of RF technology, nor is it intended to be. A common denominator among these emitters of non-ionizing RF energy however, is that all employ relatively high power levels — some reaching tens of kilowatts or more and require careful control.

A Brief History

Concern over potentially hazardous effects of non-ionizing (RF) radiation goes back many years. In 1966, the American National Standards Institute (ANSI) developed a voluntary concensus standard which recommended limiting human exposure to RF energy to a level of 10 mW/cm^2 from 10 MHz to 100 GHz and 1 mWh/cm^2 during any 1 hour period. In 1970, the Bureau of Radiological Health (now the National Center for Devices and Radiological Health) of FDA adopted manditory leakage limitation standards for microwave ovens of 1mW/cm^2 upon manufacture, and 5 mW/cm^2 thereafter, when measured at 5 cm from the surface of the oven. In 1974, ANSI revised the C95.1 standard to account for near field situations, where the electric and magnetic fields may not contain equal energy, by adding limits of 200 V/M electric field strength and 0.5 A/M magnetic field strength to the 10 mW/cm^2 power density limit. The U.S. Department of Labor adopted a 10 mW/cm^2 maximum safe exposure limit recommendation in 1975.

Based on additional research and available data, ANSI recently issued a new American National Standard: C95.1-1982 "Safety Levels with Respect to Human Exposure to Radio Frequency Electromagnetic Fields, 300 KHz to 100 GHz." This new standard differs from previous ANSI standards in several significant ways. The frequency coverage has been expanded to include 300 KHz to 10 MHz; the recommended exposure limits have been reduced, some by as much as 10:1, for all frequencies covered by previous ANSI standard; the recommended exposure limit *depends on frequency.*

By extending the lower frequency limit down to 300 KHz, ANSI now provides guidance for many types of equipment which had never been covered by an applicable standard.

Range	Density	H^2
MHz	mW/cm^2	A^2/m^2
0.3 – 3	100	2.5
3 – 30	900/f^2	0.025 (900/f^2)
30 – 300	1.0	0.025
300 – 1500	f/300	0.025 (f/300)
1500 – 100,000	5	0.125

Note: f is the frequency, in Megahertz (MHz)

Fig. 1a

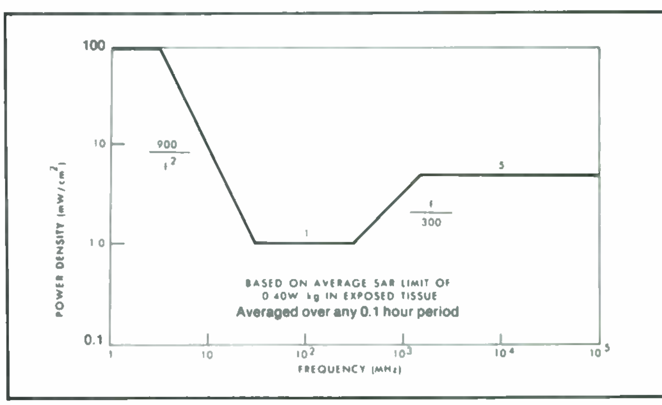

Fig. 1b

Figure 1. Radio Frequency Protection Guides:

For human exposure to electromagnetic energy at radio frequencies from 300 kHz to 100 GHz, the protection guides, in terms of the mean squared electric (E^2) and magnetic (H^2) field strengths and in terms of the equivalent plane-wave free-space power density, as a function of frequency, are given the table.

For near field exposures, the only applicable protection guides are the mean squared electric and magnetic field strengths. For convenience, these guides may be expressed as the equivalent plane wave power density.

For mixed or broadband fields at a number of frequencies for which there are different values of protection guides, the fraction of the protection guide incurred within each frequency interval should be determined, and the sum of all such fractions should not exceed unity.

In addition to thousands of AM broadcast transmitters, this frequency band includes radio navigation transmitters, various two-way radio communication bands, and equipment for induction melting, brazing, welding and tempering of metals. Also included are heating and drying devices used in laminating and veneering operations. It is not unusual to find multi-kilowatt industrial equipment in these applications, with some large scale systems running into the hundreds of kilowatts.

The new C95.1-1982 ANSI Standard includes a Radio Frequency Protection Guide (RFPG) (see figure 1 & 2) which, based on ANSI's interpretation of available literature, is designed to limit the absorption of energy by the human body. This is derived from the concept of a Specific Absorption Rate (SAR) of energy deposition of 0.4 watts per kilogram averaged spatially and temporally over the entire body mass for a 0.1 hour period.

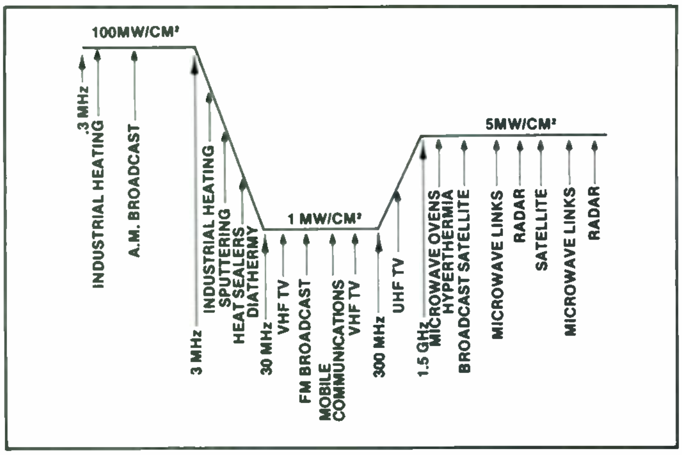

Figure 2.
The ANSI C95.1-1982 RFPG levels vary widely for the literally hundreds of thousands of emitters which operate in the frequency spectrum from 300 kHz to 100 GHz. The frequencies of operation and RFPG levels for a representative sample are shown.

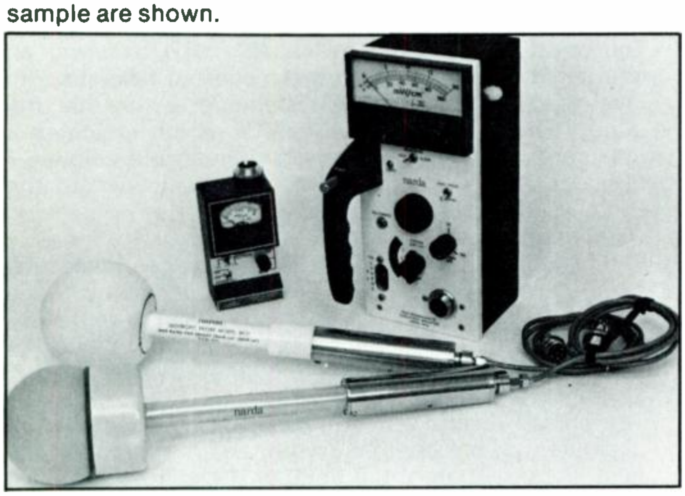

Figure 3.
Radiation Monitoring Systems offer a wide variety of metering instruments and field sensitive probes. Selection of the appropriate compliment of equipment depends on the particular application.

Since the body as a whole exhibits frequency-dependent rates of absorption of RF energy, the recommended exposure limits have been made frequency dependent in order to reduce the allowable field strength across the range of frequencies in which a human body might be expected to reach maximal absorption (whole-body resonance).

The standard also limits *peak* SAR's to 8 W/kg averaged over any *one gram* of tissue (to prevent hot spots).

Note that from 10 MHz, where the RFPG is 9 mW/cm^2, through 100 GHz the new ANSI standard is below the old 10 mW/cm^2 limit at all frequencies. From 30 MHz to 300 MHz, the limit is 1 mW/cm^2 — a reduction of 10:1. From 300 MHz to 1500 MHz the RFPG increases to 5 mW/cm^2 where it remains out to 100 GHz — a 2:1 reduction from the 1974 standard. The actual RFPG is shown in Figure 1.

Why Monitor?

Of course, the most obvious answer to the monitoring question is for operator safety. Surely, no one disagrees that gross effects which lead to arcing, burns, or overall body temperature elevation are to be avoided. While debate continues with regard to short and long term bioeffects of exposure to "moderate" levels of RF energy, a rational approach to controlling human exposure should involve accurate determination of exposure levels and evaluation of the potential risks in light of available guidelines. Once the RF levels have been determined and evaluated, the immediacy, cost and benefits of taking steps to control the level of exposure can be properly assessed.

Monitoring leads to increased safety for both the individuals likely to be exposed *and* for the organizations responsible for the design, manufacture, and operation of the RF emitter creating the exposure. Employee concern over safety can lead to reduced productivity and even work stoppages. While product liability laws are complex, one might reasonably expect that the manufacturer who makes a conscious effort to insure that his product is safe according to currently accepted standards will be in a strong position should there be a claim against him. Likewise, the employer who takes reasonable steps to create a safe working environment for his employees is likely to be viewed favorable by both his employees and by regulatory agencies such as OSHA, in addition to minimizing his own liability.

While the majority of concern over RF exposure has focused on occupational exposure, the states of New Jersey and Massachusetts, and Multnomah County, Oregon have moved to institute general public exposure guidelines.

With the Environmental Protection Agency working on what may become a manditory national public exposure regulation, and the Federal Communications Commission moving toward putting "teeth" into OSHA and EPA guidelines through its enforcement authority, now is the time to recognize the need for RF hazard control, on one's own schedule and budget, *not* when the citation is issued and the "clock is running."

That such potentially hazardous situations exist is well documented. For example, both the U.S. Bureau of Radiological Health (BRH, now NCDRH) and the Radiation Protection Bureau of Health and Welfare Canada have issued reports which include survey data citing exposures in the 100 to 300 mW/cm^2 range for operators of industrial RF

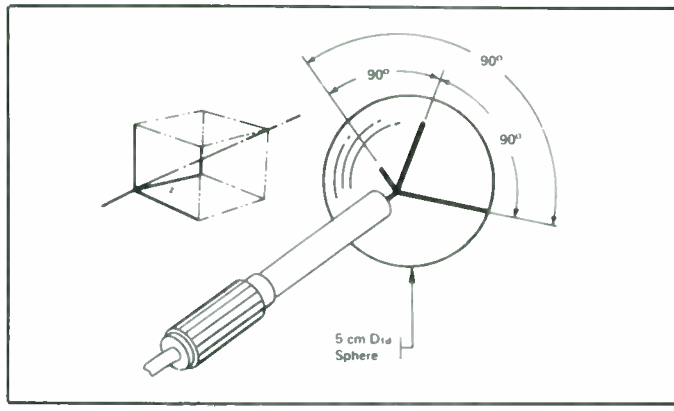

Figure 4.

Every probe consists of three mutually perpendicular lossy probe elements. Each element provides a DC output signal proportional to the square of the electric field strength incident to the element. The summation of the DC signals from the three orthogonal probe elements provides a measure of the total energy or equivalent power density, independent of direction or polarization of the RF signals. The detectors for the DC signal generation are thin-film thermocouples that provide true square-law outputs. The DC signal is proportional to the power dissipated in the thermocouple elements and the average energy density in the volume in which the elements are contained.

equipment — levels on the order of those used in diathermy treatments. Clearly, the possibility of over exposures such as these must not be overlooked.

Monitoring Protocol

Measurement of potentially hazardous fields is discussed from a theoretical standpoint in ANSI C95.5-1981 "Recommended Practice for the Measurement of Hazardous Electromagnetic Fields — RF and Microwave". In practice, one purchases a commercially available survey instrument such as that shown in figure 3.

Some of the features and functions required for actual measurements will vary depending on the particular application, but there are certain basic features which are essential.

1. Equivalent Power Density Range:
   The instrument should be capable of measuring an equivalent power density or mean squared field strength level equal to the allowable standard within that frequncy range. Thus a survey meter which indicates a maximum of 20 mW/cm^2 may be unuseable below 6.7 MHz for ANSI compliance measurements, where the RFPG value exceeds 20 mW/cm^2.
2. Dynamic Range:
   The instrument should offer a suitable range of measurement above and below the RFPG level. A dynamic range of 30 dB (1000:1) between the highest and lowest readings is typical, with the low end reaching at least 1/5 of the lowest RFPG value in the frequency range.
3. Isotropic Response:
   The survey results obtained should be essentially independent of the orientation of the survey probe in the field, the direction of the field, or its polarization. This includes the requirement that the probe's sensing element and the handle on which it is attached not perturb or distort the local fields, which would cause erroneous readings. See figure 4. A typical value for isotropic response is ± 0.5 dB maximum variation.
4. Square Law Response:

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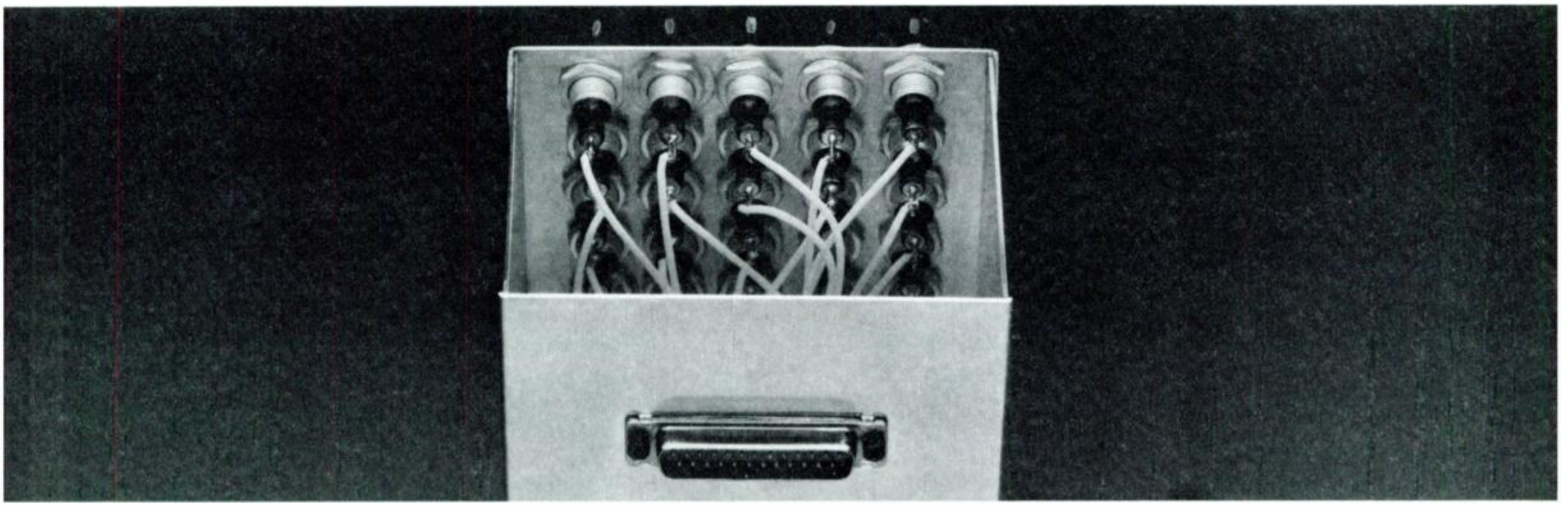

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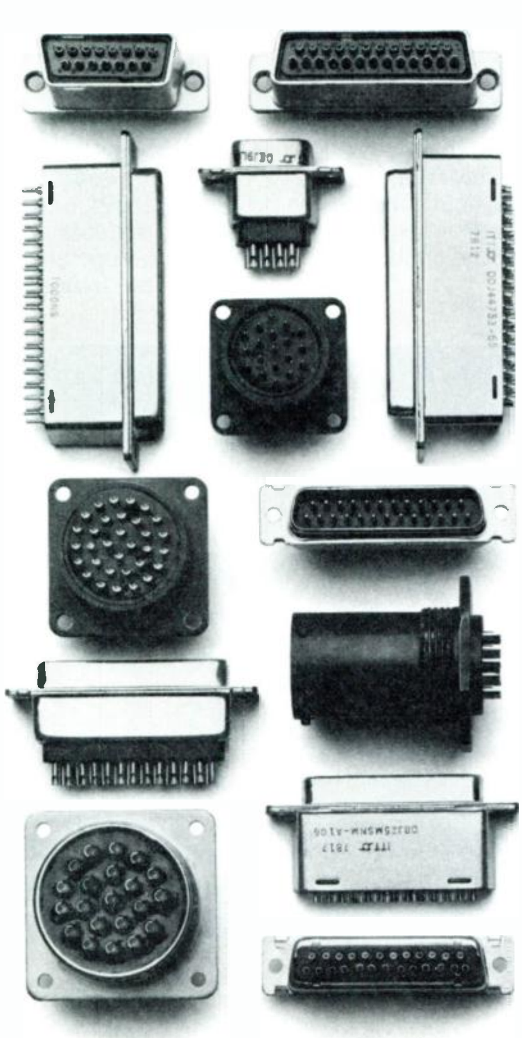

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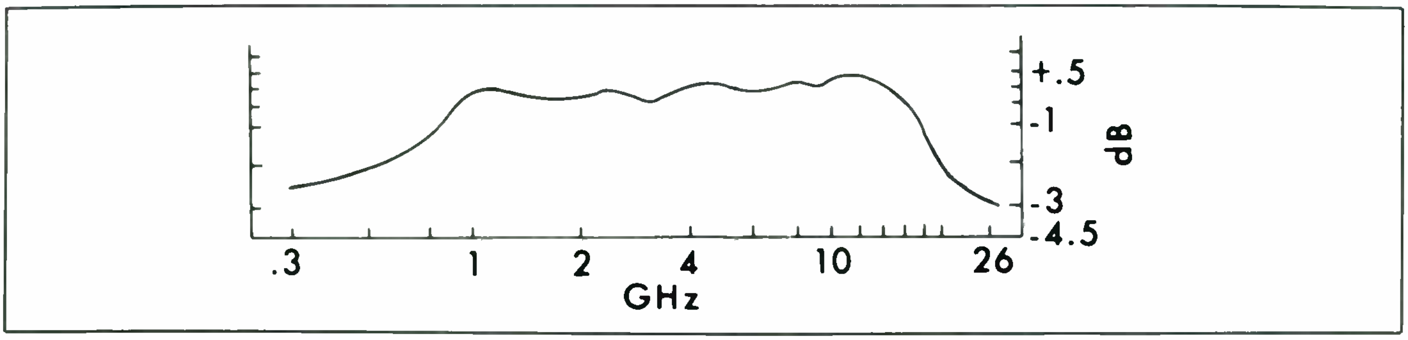

Figure 5. Typical frequency response characteristics.

Where deviation from an ideal "flat" response become appreciable, use of the correction factor permits accurate measurements independent of the actual shape of the probe characteristics.

The reading obtained should be directly proportional to the equivalent power density (mean-squared field strength) over the dynamic range of measurement. Further, the reading should be proportional to the true R.M.S. average value of the equivalent power density even if the signals are modulated, pulsed, or if multiple signals are present. An instrument which employs a peak detector cannot, in general, accurately indicate the true equivalent power density except in the special case of a single CW signal.

5. Frequency Sensitivity:
Most sensors are designed to be "flat" (Figure 5) that is, to exhibit constant sensitivity versus frequency. In practice, the meter indication obtained in a constant field may vary by as little as ±0.5 dB (+12%, -11%) or as much as ±2 dB (+58%, -37%). The uncertainty of measurement in the latter case can lead to surveys which are either unnecessarily conservative or unacceptably liberal. The solution to this problem is calibration of each individual unit at multiple frequencies across the operating range of the probe. The calibration process yields a correction factor equal to the reciprocal of the deviation of the sensitivity from the ideal response. The meter reading, when multiplied by the correction factor for that frequency, yields the true value of the field strength to within ±0.5 dB. (Incidentally, a ±0.5 dB uncertainty is on the order of the uncertainty of a probe calibration by the National Bureau of Standards).
6. Electrical and Magnetic Field Capability:
In near-field monitoring situations, below approximately 300 MHz, "reactive" fields may be present, causing unequal distribution of energy between the electric and magnetic fields. Thus for near field exposures it is necessary to measure the mean-squared electric and magnetic fields separately. Note that the near field region may extend from ½ to several wavelengths from an emitter. (In the far field, it is sufficient to measure either field, as they are related by the impedance of free space $H^2 \times (377)^2 = E^2$).

Additional features which should be considered in the selection of a survey instrument are portability, (size, weight, transit case), ease of use (zero adjustment, scale markings, maximum hold mode, chart recorder output), power supply (rechargeable or disposable batteries and/or line voltage operation), pulse and CW overload and burn-out levels, audible alarm to warn the surveyor of excessive field levels, and availability of periodic (annual) recalibration service.

Multiple Signals

The previous ANSI standard, being constant with frequency, tended to minimize the difficulties encountered in making multiple frequency measurements. If the sensor had relatively constant (flat) sensitivity to all of the signals, then the meter indication produced by a square law sensor

(Continued on page 20.)

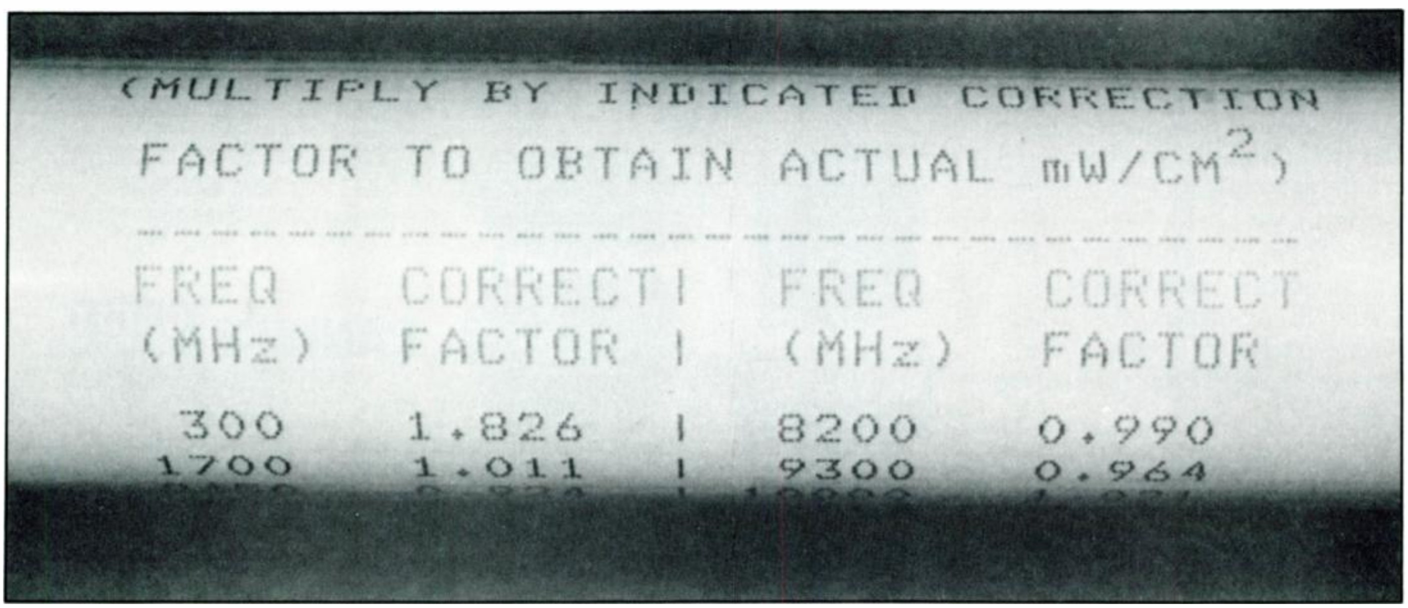

Figure 6.

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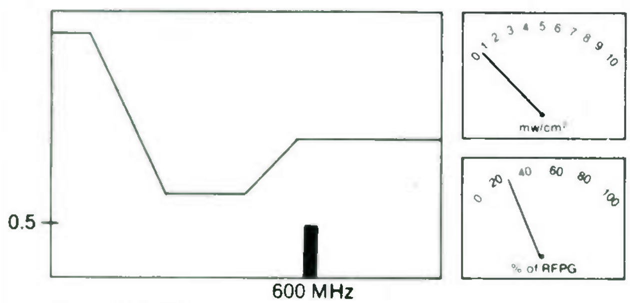
INFO/CARD 11

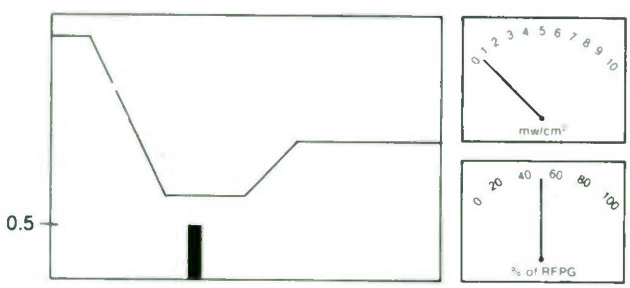
Application Of New ANSI Standard In Multiple Signal Situations.

For example, the RFPG for television broadcast varies depending on the band. VHF TV is within the 30-300 MHz region (1 mW), while the RFPG for UHF TV varies from 1.57 mW at 470 MHz to 2.97 mW at 890 MHz. For a hypothetical UHF transmitter creating a field of 0.5 mW/sq. cm. at 600 MHz, a "flat" R.M.S. reading probe would measure the true equivalent power density level, corresponding to 25% of the RFPG; a peak detecting probe would read roughly 0.5mW/sq. cm. depending on the error (perhaps as much as +/− 40%) introduced by the modulation of the TV carrier; and a shaped probe would read 25% of the RFPG directly.



(A) UHF TV

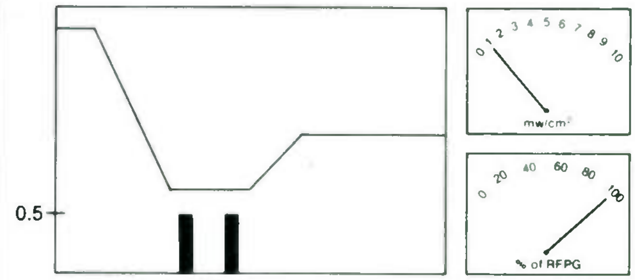
Likewise, for a VHF TV transmitter creating a 0.5 mW/sq. cm. field, the readings would be 0.5 mW/sq. cm, 0.5 mW/sq. cm (+/− 40%), and 50% of the RFPG for the three sensor types respectively.



(B) VHF TV

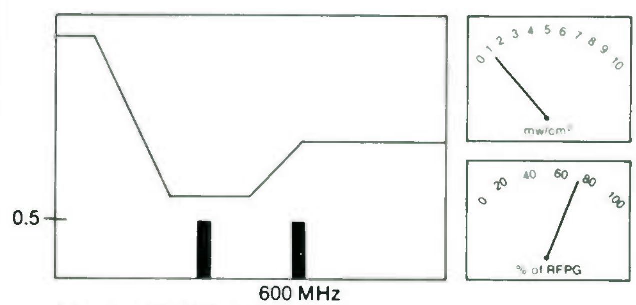
Two VHF transmitters with the same output power and field strength level (0.5 mW/cm. sq. each) would cause the square law (R.M.S.) sensor to indicate the sum of the two power levels, or 1.0 mW/sq. cm. The shaped probe would also indicate twice the exposure level, or 100%. A peak detector, however, could indicate as high as 2.0 mW/sq. cm. or as low as 0.5 mW/sq. cm., depending on the exact interactions of the signals (plus an additional error due to the modulation of the signals).

If VHF and UHF transmitters share the same tower, with each generating a 0.5 mW/sq. cm. field at the point of measurement, the flat R.M.S. reading probe will respond to the true net equivalent power density and read 0.5 + 0.5 = 1.0 mW/sq. cm. (The peak detecting probe could still read as much as 2 mW/sq. cm. or as little as 0.5 mW/sq. cm.). How should this reading be interpreted? According to the C95.1



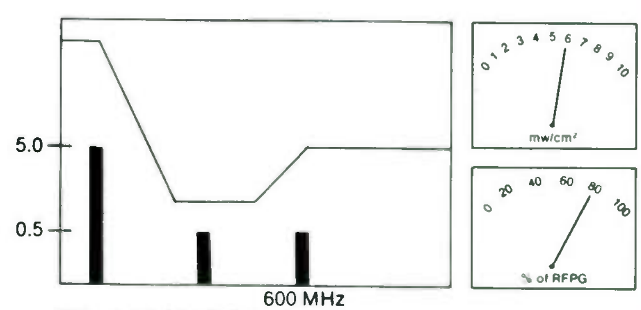
(C) TWO VHF TV

Standard, for multiple signals "the fraction of the protection guide incurred within each frequency interval should be determined, and the sum of all such fractions should not exceed unity". In fact, the shaped sensor would "normalize" each signal as a percent-of-standard independently, before adding their contributions, and would indicate 75% of the RFPG, the correct value.



(D) 1—VHF & 1 UHF

To complicate the situation further, now add an AM broadcast transmitter which contributes a field of 5 mW/sq. cm. The square law "flat" probe would indicate 5 + 0.5 + 0.5 = 6 mW/sq. cm., essentially obsuring the VHF and UHF signals. The response of a peak detector in this situation is almost impossible to predict. The shaped probe however, will add the contribution of a signal which is only 5% of the RFPG at that frequency and show the correct value of 80% of the RFPG.



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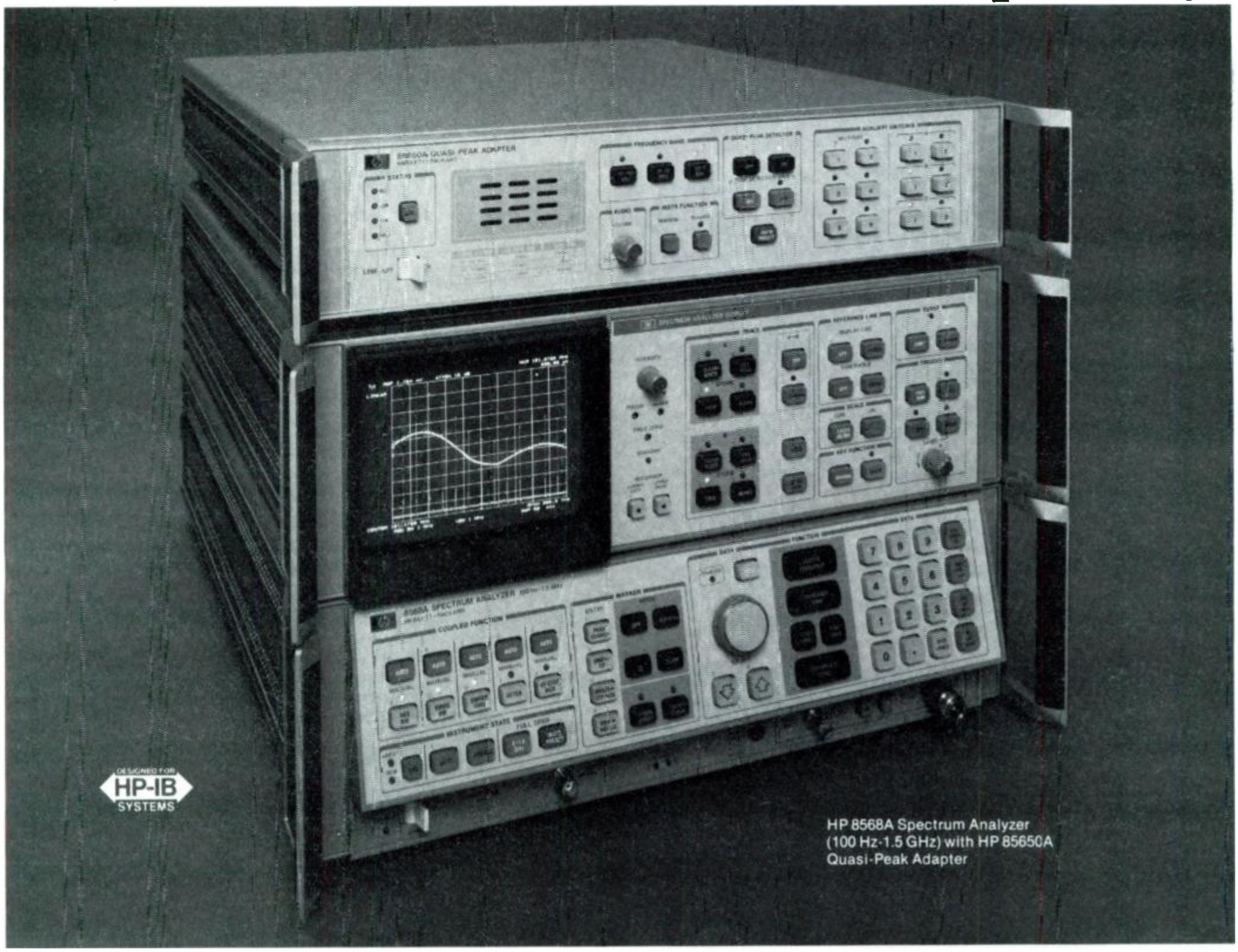

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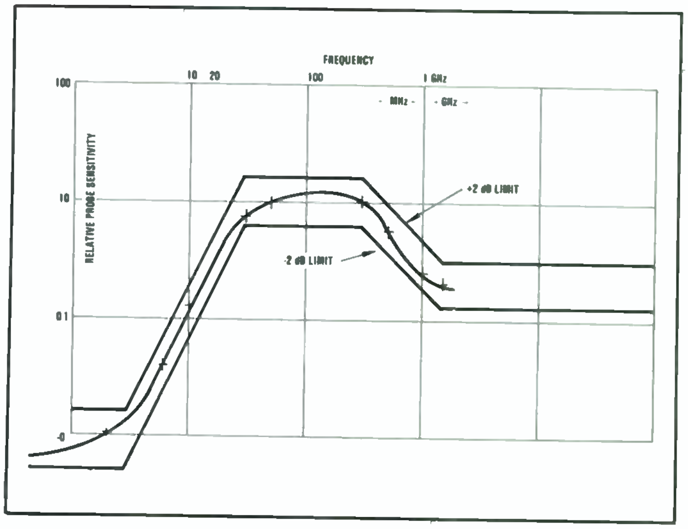


Figure 7.

The relative probe sensitivity of the Model 8682 ANSI shaped probe is such that all signals which are at levels equal to the same *percentage of the RFPG* for their frequency produce equal output signals.

would be proportional to the net equivalent power density and thus the net exposure level was easily computed. In order to employ a "flat" sensor in a situation where various signals are present, at frequencies where the RFPG levels are substantially different, the measurement problem becomes complex. If each of these signals cannot be measured independently, a spectrum analyzer might be required to determine their relative amplitudes, compounding the implict requirement for a priori frequency information in order to determine the appropriate value of the RFPG in the first place. There is, however, another way.

Instead of a probe which exhibits a constant output or sensitivity versus frequency in a constant intensity field, consider a probe which responds with a constant output versus frequency in a field which is a *constant fraction of the ANSI RFPG* (see figure 7).

The frequency sensitivity of such a probe would have to mirror the ANSI standard over the range where the RFPG varies by 20 dB (100:1).

Such a probe, the Narda Model 8682, is currently being manufactured. The resulting meter readings indicate directly the *percent of E-field exposure level according to the C95.1-1982 RFPG,* vastly simplifying the measurement process.

While the Model 8682 operates from 300 KHz to 1500 MHz, covering the entire variable portion of the ANSI RFPG (which is also the spectrum where most multiple signal situations are likely to occur), the U.S. Navy has commissioned Narda to develop an ultra broadband shaped frequency response radiation hazard test set covering 2 MHz to 40 GHz with a dynamic range from 3% of the ANSI RFPG to an intermittent exposure level equal to 3000% of the ANSI RFPG. This will lead to the development of a commercial unit which will cover 300 KHz to 26 GHz with a dynamic range from 1% to 1000% of the RFPG.

Typical Systems

For near field measurements (below 300 MHz) a probe such as the Model 8662 will measure the Electric field from 300 KHz to 300 MHz from 0.2 up to 200 mW/cm^2. This should be combined with one or two magnetic field probes such as the 8652 (300 KHz to 10 MHz; 0.2 to 200 mW/cm^2) and 8631/

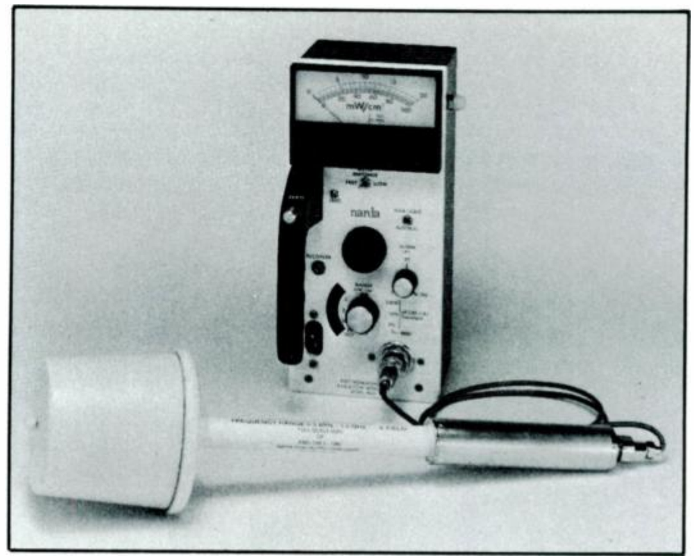


Figure 8.

The Model 8682 ANSI shaped probe is compatible with existing metering instruments with the addition of scale markings for 0-2%, 0-20% and 0-200% of the RFPG.

8633 probes (.02 to 20 mW/cm^2 and 0.1 to 100 mW/cm^2 respectively; 10 MHz-300 MHz). These probes all have flat frequency response; the Model 8682 ANSI shaped probe may be substituted for the Model 8662. Selection of the metering unit for use with the system depends on the features and functions required.

Far field measurements can be made with the 8682 probe from 300 MHz to 1500 MHz. For higher frequencies, probes as the 8621/8623 (.02 to 20 mW/cm^2 and 0.1 to 100 mW/cm^2 respectively) cover 300 MHz to 26 GHz with calibrations available on request up to 40 GHz. All probes and meters are interchangeable.

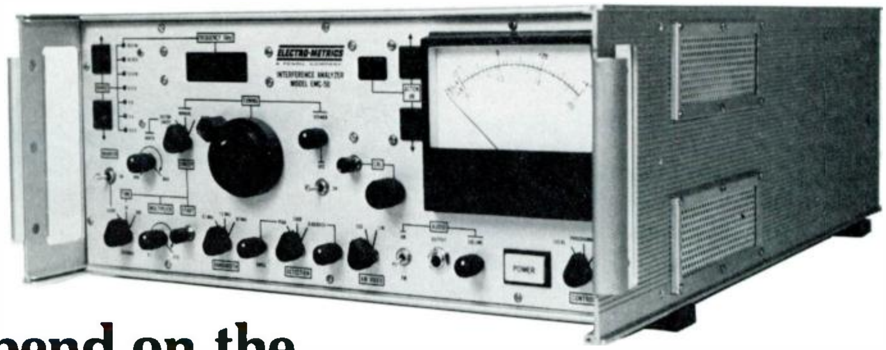
Summary

Both the awareness of the potential hazards to health and safety from exposure to RF energy, and instrumentation for the measurement and control of these fields have natured considerably in the last two decades. The current consensus standard (C95.1-1982) recognizes the physical phenomena which impact exposure level by providing a frequency dependent Radio Frequency Protection Guide, and by explicitly requiring the separate determination of both electric and magnetic field strength in near field situations. A large number of emitters which had previously been exempt from any standard have now been included in the expanded frequency range covered by the C95.1-1982 standard.

Instrumentation for use in commercial and industrial applications should be selected with due consideration for accuracy, frequency range, power level, and ease of operation. Below 300 MHz, separate electric and magnetic field measurements should be made. Above 300 MHz, electric field measurements alone are normally adequate.

Particular attention must be paid to multiple signal situations where the signals occupy frequency intervals with different RFPG levels. Fortunately, the technology exists to fabricate sensors which automatically incorporate the frequency dependence of the RFPG into their readings, making today's instrumentation readily useable by operators of all levels of expertise, and thereby putting non-ionizing radiation hazard measurement and control within reach of all concerned individuals and organizations.

10 Good Reasons



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Prediction of EMI Radiation From PCBs

By Michel Mardiguian
Don White Consultants, Inc.
Gainesville, VA 22065

Background

Back when radio frequency interference was primarily a matter of radio communications transmitters and receivers, very sophisticated analytical approaches were developed to predict the interference level knowing all the parameters of the transmitting source. Then, a multitude of non-intended RF sources, producing discrete frequencies, have appeared with the proliferation of computing devices using faster and faster logic technologies to allow corresponding faster clock and bit rates. All computers and digital equipment therefore, became a threat for radio communication. A simple calculation can give an immediate feeling of the problem — let us assume a mini-computer consisting of a large planar board with 60 chips, each one consuming about 250 mwatts. Assuming also that only one-third of the circuits resident on these chips are toggling synchronously at an internal clock frequency of 50 MHz, for instance. We can say that the total switched power at a given instant is:

$$1/3 \times 60 \times 0.250 = 5 \text{ watts.}$$

Let us now assume that a miniscule fraction of this power is *not dissipated* by Joule effect in the chips, the wiring and the various resistances, but is radiated instead. And, let us assume that on the 50 MHz fundamental, only 10^{-6} switched power is radiated. One-millionth of 5 watts is 5 microwatts. A simple formula gives the field strength of a given radiator:

$$E \text{ volts/m} = 1/R\sqrt{30 P_r} \qquad (1)$$

where: R = distance from the source in meters
P_r = radiated power (including antenna gain)

Figure 1. Predicted E field levels at 1 meter distance from a unity (1 volt, 1 cm²) radiating circuit.

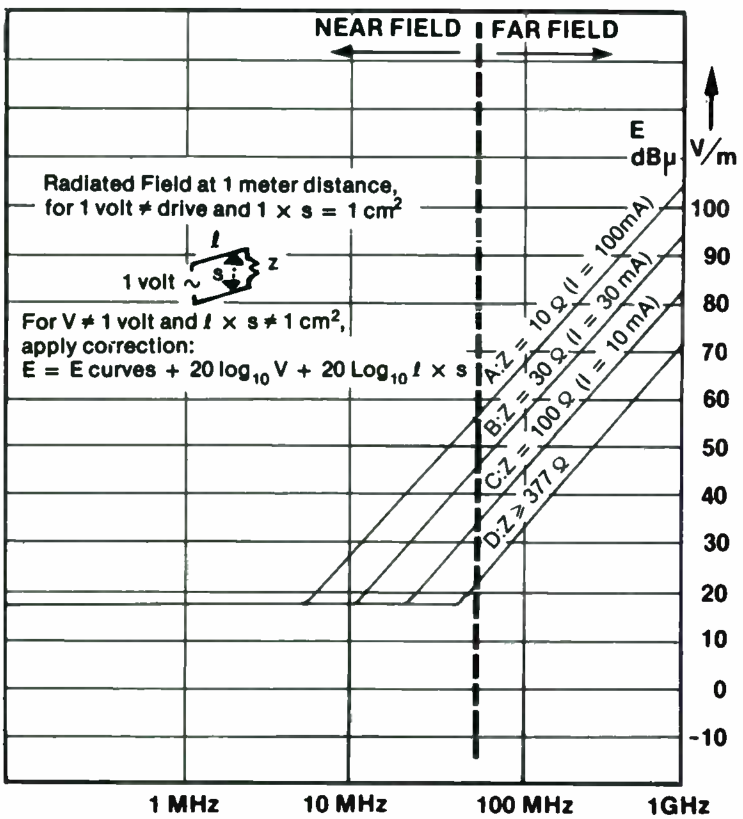
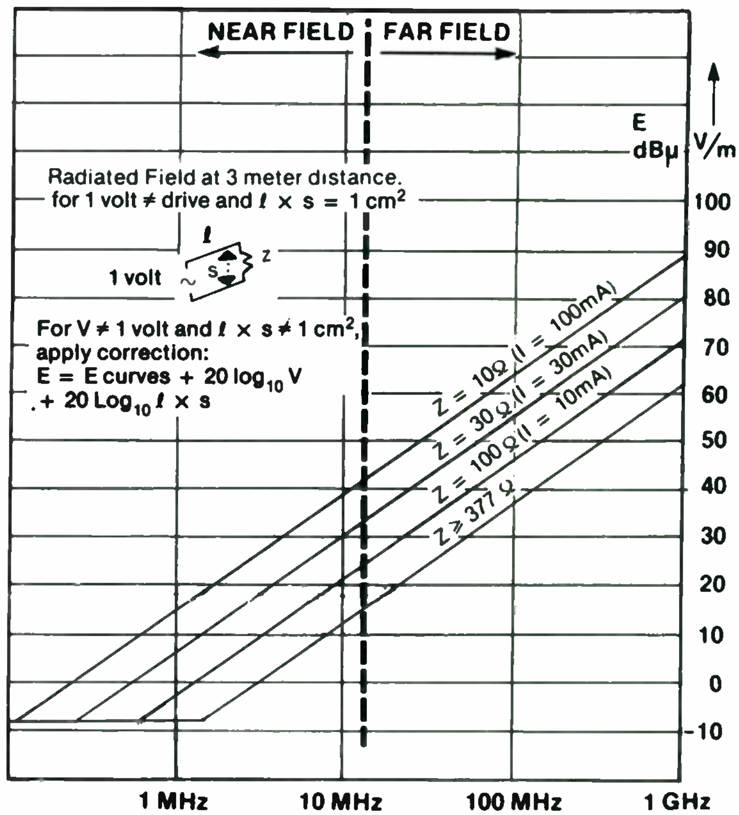


Figure 2. Predicted E fields levels at 3 meter distance from a unity (1 volt, 1 cm²) radiating circuit.



Predict the interference level that can be expected from the near proximity of a computer or digital circuit on your communication systems.

At three (3) meters distance, our five (5) microwatts radiated from the PC board will result in:

$$E = 1/3\sqrt{305 \times 10^{-6}} = 4\text{m Volts/m or } 72\ \text{dB}\mu\text{V/m}$$

This is more than the minimum field strength received by TV and FM listeners in areas remote from the broadcast station. Therefore, in case of frequencies coincidence (co-channel EMI), the computer clock and its harmonics may seriously affect radio-reception in the vicinity.

The FCC and other regulatory bodies world-wide have set RF emissions limits, like FCC Docket 15J which stipulates no more than 100 μV/m (40 dBμ/m) measured at 3m distance from personal computers, above 30 MHz. But to the contrary of a CW Transmitter where the characteristics of the radiation source (transmitter power, antenna gain and pattern, spurious harmonics, etc.) are well identified, a digital electronic assembly is much more difficult to model. So the traditional approach is generally to design the PCB layout, mother board and interconnects using the best know-how (which, in the area of EMI may include a blend of rules-of-thumb, company's recipes and "we-do-it-this-way-because-we-always-did-it-this-way-and-it-worked"), then to run an FCC test to "see if it passes". Needless to say (and the example above shows why), in many cases, "it does not pass", which means re-design, E.C.'s and retrofits.

The following method allows quick prediction of radiation from printed circuits and associated wiring.

The prediction is based on solutions of Maxwell's equations for a small loop or doublet. The far field electric term from a short doublet is given by:

$$E\mu V/m = \frac{I\,\ell\,60\,\pi}{R\,\lambda} \qquad (2)$$

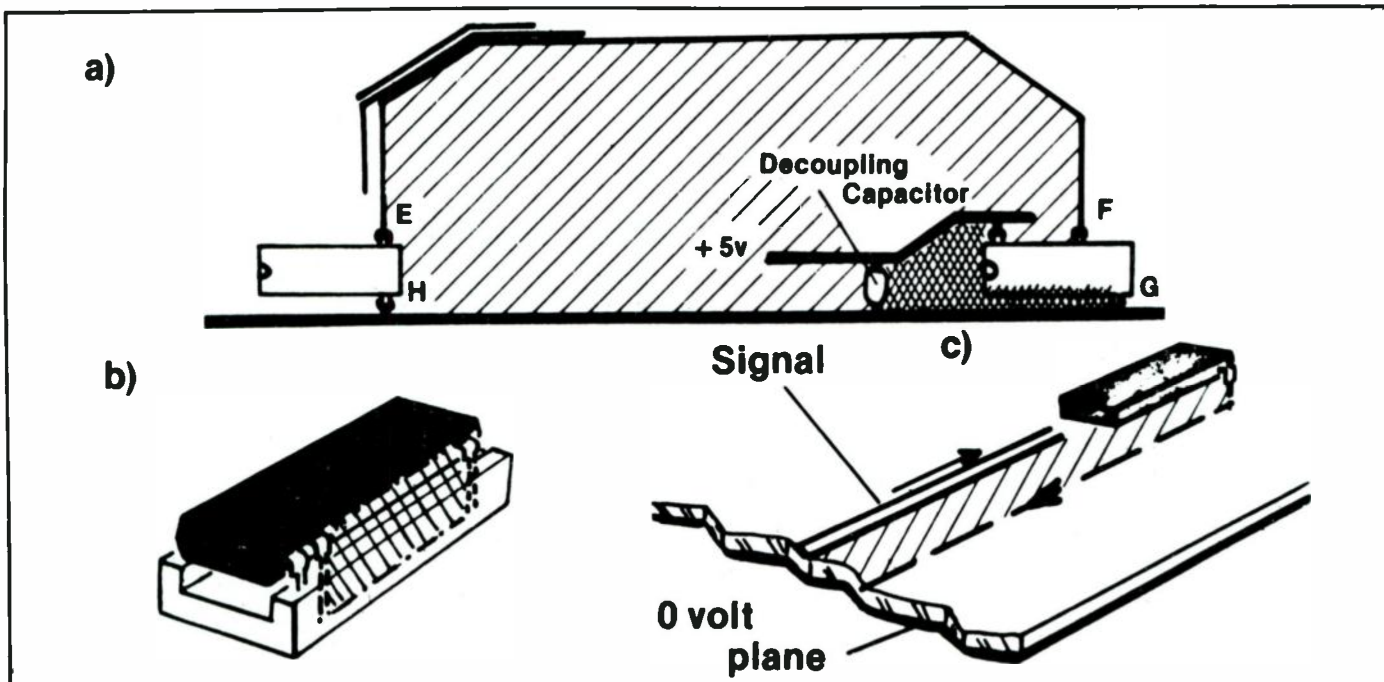


Figure 3. (a) Radiating loop formed by supply traces (between the decoupling capacitor and the chip), and by a signal trace and its return.

(b) Radiating loop formed by the chip and socket leads; with 50 chips or more on a PC Board, the sum of these small areas can represent up to 30 - 50 cm², a non-negligible contribution to total radiation!

(c) The radiating loop is immediately reduced when the return current can flow in a ground plane just under the trace.

where: I = current in the short wire, in Amperes
ℓ = length of short wire in meters
R = distance in meters
λ = wavelength in meters

Since our radiating circuit is made of two wires more or less parallel carrying opposite currents, the net field would be null if they were infinitely close, but in fact, their separation "S" creates a phase difference, therefore, a non-null net field. Taking into account this phase shift gives the radiated far-field expression for a pair of parallel wires:

$$E\ V/M = \left(\frac{I\,\ell\,60\pi}{R\lambda}\right)\left(\frac{\sin 2\pi S}{\lambda}\right) \cong \frac{I\,\ell\,120\pi^2}{R}\left(\frac{S}{\lambda^2}\right) \quad (3)$$

Since the sine of a small value approximate this value itself.

Given that $\lambda m = \frac{300}{F(MHz)}$ replacing $\ell \times s$ by the area of the circuit, and using more convenient units results in:

$$E\mu V/m = 1.3 \times A \times I/R \times (F)^2 \quad (4)$$

where: A = area of the radiating circuit in cm^2
F = frequency in MHz

Interestingly, the result is the same whenever we use a loop or two parallel wires.

However, these equations are based on current, which is generally not what the designer uses the most. It is often more convenient to use the drive voltage. When I is replaced by V/Z in equations, the following equation results, which expresses directly the radiated field as a function of the driving voltage:

$$E\mu V/m = \frac{1.3\,A\,V\,(F)^2}{R\,Z} \qquad (5)$$

The curves Figure 1 and 2 give the field strength radiated at one meter and three meters from a circuit normalized as a 1 cm^2 loop area, driven by 1 volt, terminated by loads from 10 ohms to 377 ohms. Above the far-field transition, if loads are larger than 377Ω, the field becomes essentially of a voltage-drive dipole, i.e., it relates to the driving voltage only. What remains to be done is to compute (or measure) the Fourier series of the digital signal.

The results are fairly accurate as long as the radiating element is electrically short ($\ell < \lambda$). Above this, limitations are:

- at highest frequencies where the longest PCB trace exceeds $\lambda/2$, the geometry of the loop entered in the model must be clamped to $\lambda/2$ of the frequency used in calculation.

Above this, limitations are: — at highest frequencies where the longest PCB trace exceeds $\lambda/2$, the geometry of the loop entered in the model must be clamped to $\lambda/2$ of the frequency used in calculation.

– above the same boundary, the load impedance has to be replaced by the characteris impedance of the PCB traces.

The radiating loops can consist of: (See Figure 3)

– The +Vdc and return traces from the power busses to the chip. In this case, the radiation is caused by the logic transition. (See Figure 3 a)

– The chip and socket leads. (See Figure 3b)

– The signal run between chips, the path being closed via the corresponding 0 volt return trace. This is one of most critical radiating loops, since, in some cards, it can represent a significant part of the card perimeter. In Figure 3 a, for instance, the run EFGH reveals a very large loop.

– The signal and return wires in a flat cable, etc.

If the PCB is multi-layer or double-sided with ground plane, an immediate reduction of the radiating loop occurs since the spacing "S" now is merely the height of the traces above the ground plane. (See Figure 3c)

Following is a practical example showing the use of the method:

Example

Assume a 4-volt TTL signal with a 10 MHz switching rate, 50% duty cycle and 10 nsec rise time. The signal is carried on a single sided back-plane, the traces run representing a 40 cm × 12 cm (16" × 15") loop, terminated by a 500Ω load. Does this design meet the FCC-15J limits for Class B devices? Table 1 recapitulates the "envelope" calculation (see Figure 4) of the voltage spectrum, the factors to apply versus the curve Figure 3, i.e., the voltage correction in dB, and the area correction which is 20 Log A cm^2.

If several loops were switching identical loads at a synchronous frequency, the calculation should sum up all these identical loops, since they would appear like a single localized source viewed by the measuring antenna.

The prediction indicates clearly that the FCC limit is violated by 12 - 15 dB, i.e., action has to be taken to reduce the traces loop, use a ground plane or shield the equipment housing.

By comparison, Figure 4 shows the actual measurements of this PCB (the tested card uses 7404 TTL inverters)

IEEE-488

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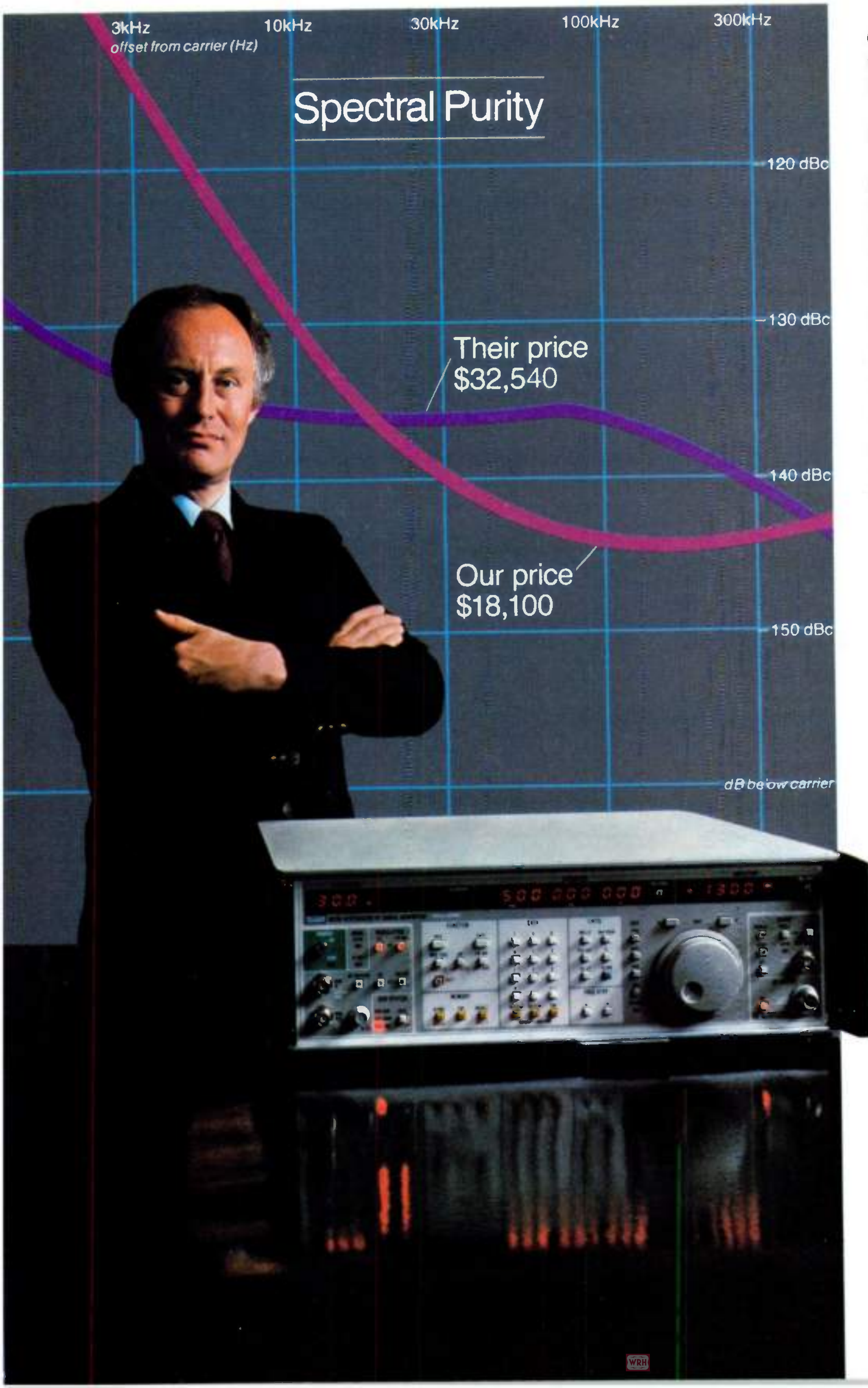

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		20 Log F1/Fx	40 Log F2/Fx				
$F_1 = 1/\pi T = 6.2$	12 dB Volt			NA	NA	54 dB	NA
Fundamental $F_o = 10$	X	−4 dB		8 dBV	−10	—	52 dBμV/m
Harmonic #2 = 20	X	−10 dB		2 dBV	−10	—	46 dBμV/m
etc.							
$F_2 = 1/\pi Tr \cong 30$	X	−14		−2 dBV	0	—	52 dBμV/m
Harmonic #5 = 50	X		−18	−6 dBV	8	—	54 dBμV/m

Harmonic #15 = 150	X		−34	−22 dBV	26	—	58 dBμV/m
etc.							

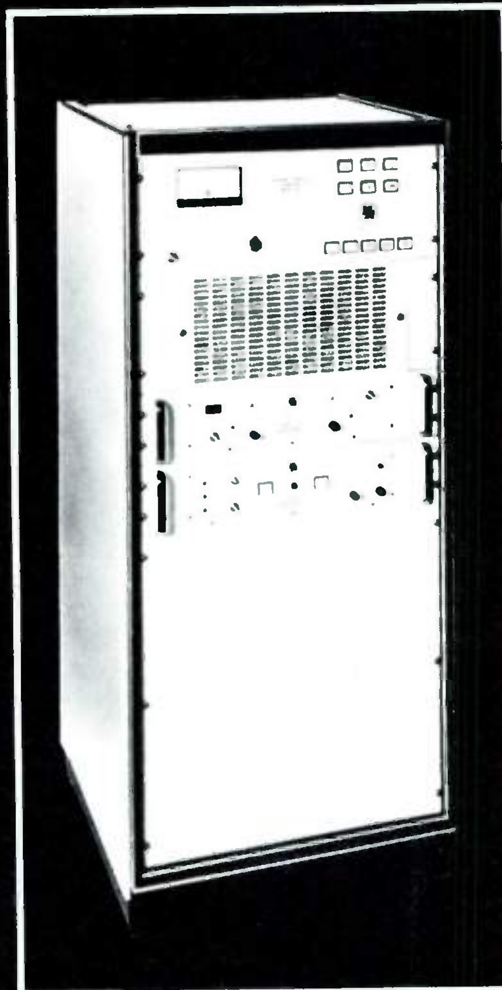
Table 1

on a 3-meters FCC test site. For sake of precision, it must be remarked that:

– The Fourier series gives peak value for each harmonic while the EMI receiver is scaled in RMS, which means −3 dB difference.

– The FCC procedure calls for scanning antenna height to search for maximum reading, which means up to +6 dB add-up when the ground reflected wave arrives in phase with the direct one.

– In an actual equipment, the radiation from common-mode currents leaking by the external cables (I/Os, etc.) may be a significant contributor to the total radiation. This path is more complicated to model, but has been studied as well and will be the subject of a future article.



Acknowledgements

The author expresses his thanks to Donald R.J. White for his tutorial support, and to Richard Harrison who designed a self-powered PCB mock-up, and to Craig MacAllister, of Carl T. Jones Associates (7901 Yarnwood Court, Springfield, VA 22153), who performed the formal test and data reduction at their FCC test facility.

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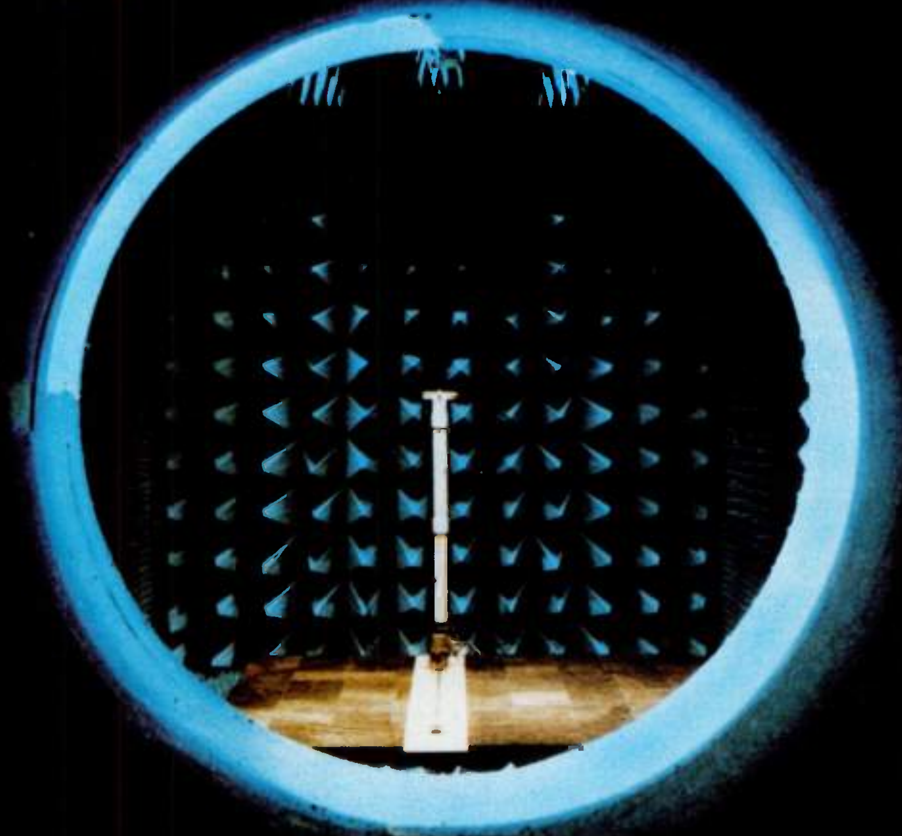
1. Donald R.J. White, *EMI Control in the Design of Printed Circuit Boards,* Don White Consultants, Inc., P.O. Box D, Gainesville, VA 22065
2. *FCC Rules and Regulations* — Part 15J
3. *Electro Magnetic Shielding Materials,* Don White Consultants, Inc., P.O. Box D, Gainesville, VA 22065

The Author

Michel Mardiguian is a Graduate Electro-Mechanical Engineer. He started in Electronic Packaging of Aircrafts, then in Computers Packaging at IBM R & D Laboratory. He entered the EMC area in 1974 and enjoys design and field experience of problems in the computer/telecommunications area. For four years he has been an active member of the CISPR (EMC studying body of the IEC) group on computer interference. He is Director of Training and Vice President of Engineering at Don White Consultants, Inc., a well-known firm which conducts seminars and consulting on Electro Magnetic Compatibility.

Looking into an Anechoic Chamber?



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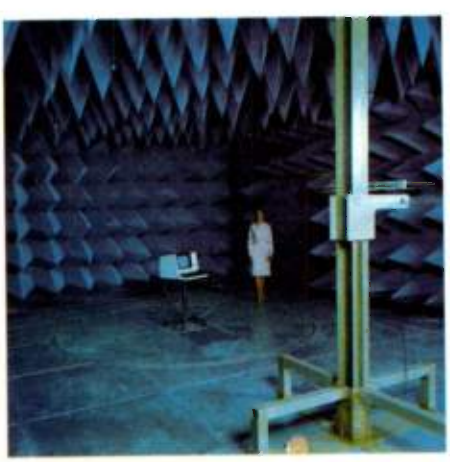

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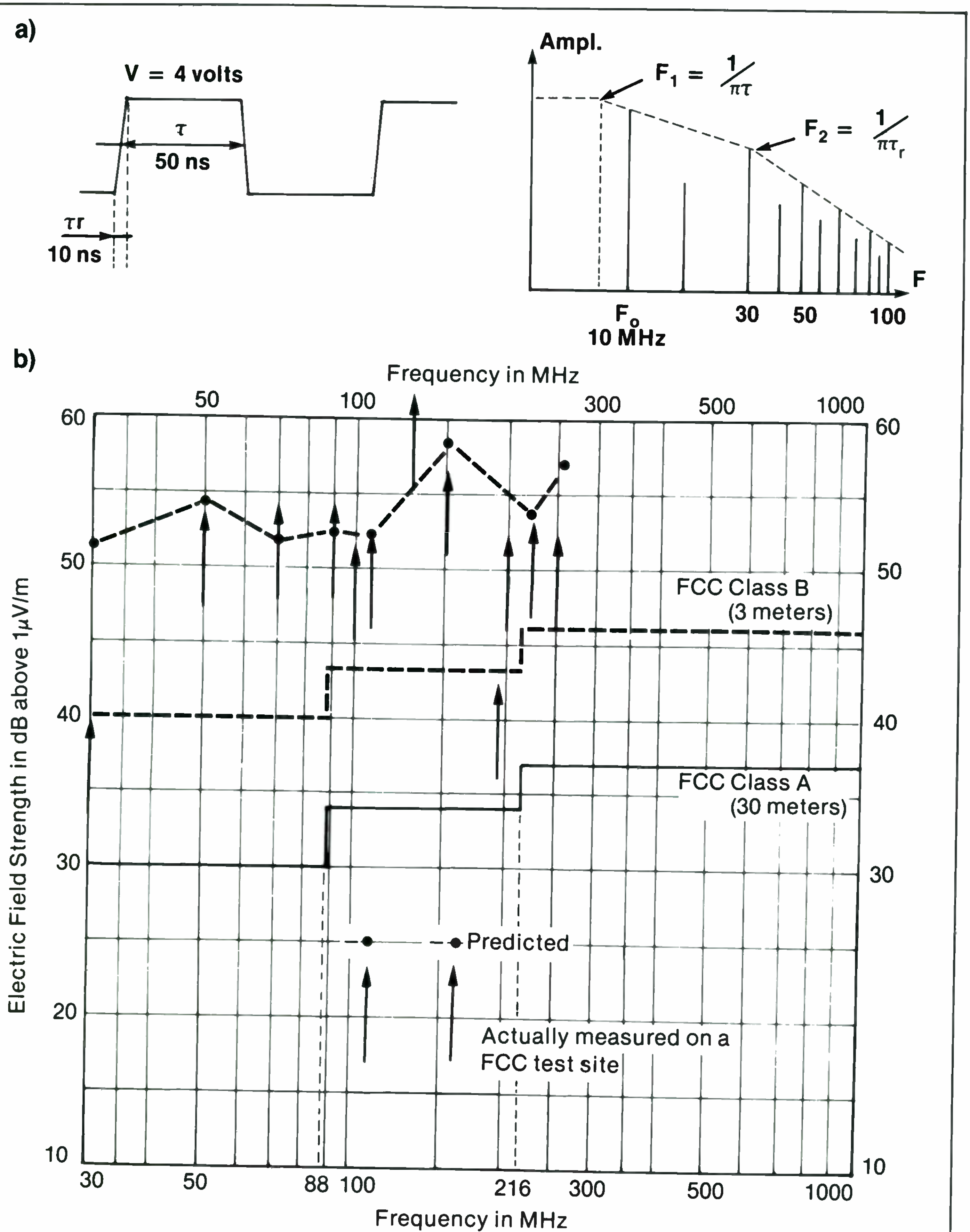

Figure 4. (a) Time and Frequency Aspects of the 4 volts pulse train.
(b) Predicted vs measured radiated field from a large PCB with 10 MHz clock.

Ferrite Beads, The Noise Fighters

Ferrite Beads offer a simple, inexpensive EMI suppression technique.

By O.V. Gigliotti
Ferrite Division
The Stackpole Corporation
St. Marys, PA 15857

In the early 1960's, ferrites formed into small sleeve configurations became known as "beads". Some typical bead configurations are shown in Figure 1. They suppress transient spikes by acting as a circuit component with high impedance to radio frequencies, yet they allow lower-frequency current to flow through the conductor without dc losses and with the introduction of very little inductance into a circuit. Use is simple and convenient — the conductor is threaded through the inside diameter of a bead. Millions are now in use, suppressing all types of transients and interfering signals (EMI) containing frequencies of approximately 1 MHz and above. They are becoming even more popular due to the FCC regulations on noise which become effective on October 1, 1983.

Some application examples include: discoidal filters using interspaced ferrite and capacitor elements; filters that use a bypass capacitor and a bead as a compact decoupling device; multi-holed, broad-band, radio-frequency ferrite chokes; radio-frequency attenuators for shielding high-explosive detonators; and for preventing activation of electronic, anti-skid brake devices, as shown in Figure 2.

Ferrite beads can be produced in a variety of electromagnetic materials, consisting of mixtures of iron oxide and metallic oxides of nickel, zinc, manganese or combinations thereof. These are calcined, milled, spray-dryed, molded or extruded and sintered at temperatures of 1100° C or higher. The resultant ferrite is a polycrystalline, ceramic material with a spinel structure.

The principle of how these ferromagnetic cores work is based on the fact that all ferrites have a permeability and quality factor ($Q = X_L/R_S$) which is frequency sensitive. Over the specific frequency range for which the material is designed, its permeability (directly proportional to inductance) and the series losses of material, R_S, are relatively constant. But as frequency increases above the operational range, permeability decreases while losses increase rather drastically (see Figure 3). Both characteristics continue in this manner until, respectively, a minima and maxima are reached. Thus we have, in essence, a frequency-sensitive "resistor" — in actual operation an impedance consisting of a decreasing inductive reactance and increasing series resistance. Each material has a frequency at which it becomes effective as a suppressor of high frequencies.

Ferrite beads have varying shapes and dimensions, depending upon their end use. The most general use of the common tubular ferrite bead is in the suppression of EMI/RFI coming into or leaving through power supply leads in high frequency equipment. Placement is very important. Beads should either be near the noise source, if known, or just outside or inside the "black box" to be protected. If positioned somewhere on a long line, they will not suppress conducted noise which has been radiated into that line on either side of it.

It is best to treat these cores as impedances. Normal filters attenuate signals by mismatching the impedances. Normal filters attenuate signals by mismatching the impedances of load and source. Most ferrite beads have relatively low impedances. Although there are MIL filter specifications, they are normally based on measurements using 50-ohm source and load impedances. In many cases, neither is known. A problem exists which is seen as a spike on an oscilloscope, for example, or jitter in the output of a black box. It is "visible" in some manner.

How much impedance is needed to eliminate this spurious signal? Robert B. Cowdell's paper[2] presents a useful formula for calculating Insertion Loss Ration providing the source and load impedances are known.

$$\text{ILR (in db)} = 20 \log \frac{Z_G + Z_L}{Z_G + Z_L + Z_M}$$

Where: Z_G = source impedance
Z_L = load impedance
Z_M = bead impedance

However, the cure in 99% of these applications is empirical. Cut and try.

(Remember, where to put it is not the question!) Placing a core into the circuit will almost always provide an answer, if not direction. A partial reduction in noise means a larger amount of material is needed. Total suppression suggests a lesser amount of material (smaller unit) may work. No effect at all indicates improper choice of material.

Material Selection

For same-size ferrite pieces of three different materials, Figure 4 compares impedance (Z) over the frequency range of 1.0 to 100.0 MHz.

Ceramag® 24 is a manganese-zinc material with a nominal permeability of 2500. Its useful suppression range is 1.0 MHz to approximately 10-20 MHz, depending on the mechanical dimensions of the part. If too large, standing waves within the bead cause a dimensional resonance effect.

Ceramag 7D is a nickel-zinc material with a permeability of about 850. It is useful from 10 to 200-300 MHz. It also has a dimensional resonance effect but in the GHz range.

Ceramag 11, another nickel-zinc material, has a permeability of approximately 125 and a range from 100 MHz up.

Beads of the first two materials are guaranteed to minimum Z values at two specific frequencies.[1] The third is tested for permeability by measuring inductance, which tends to insure its high-frequency suppression character-

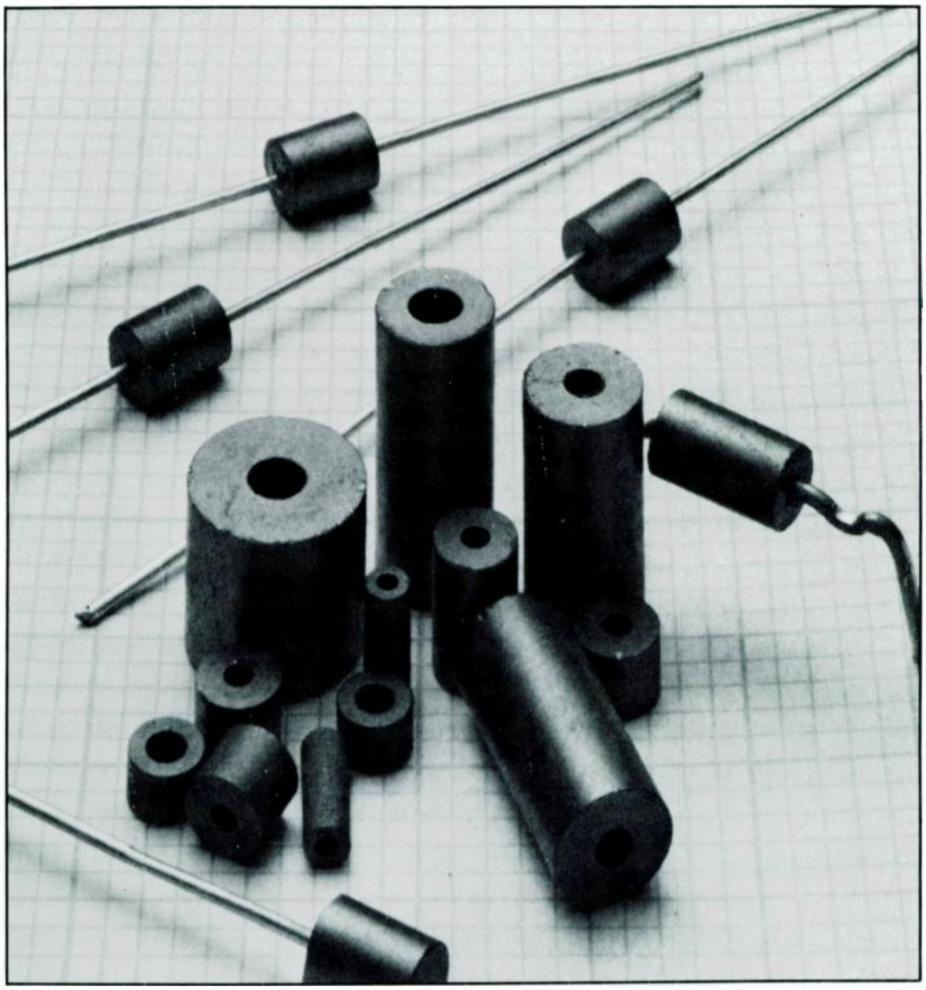

Figure 1. Small ferrite sleeve configurations known as "beads" can suppress transient spikes.

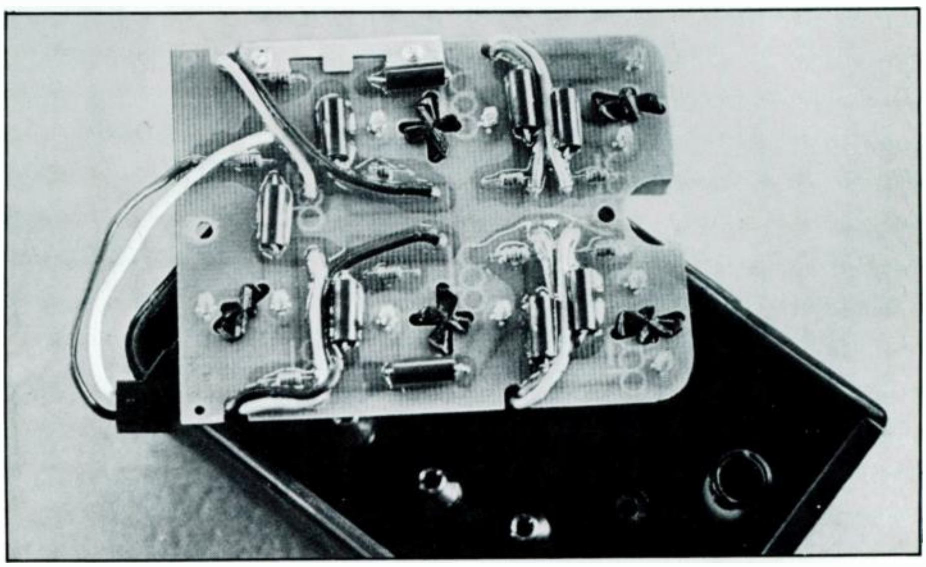

Figure 2. Nine noise suppression beads on this printed circuit board shiled anti-skid brakes from EMI.

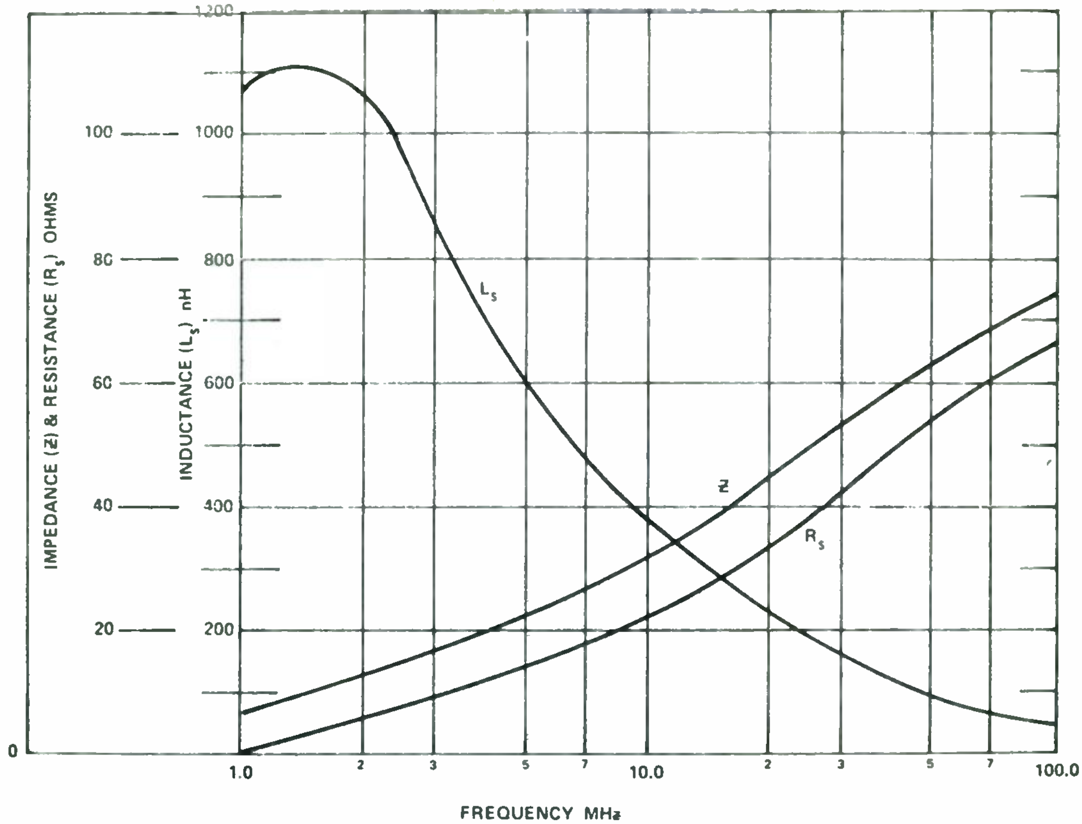

Figure 3. Typical Inductance, Resistance and Impedance vs. Frequency curves of a ferrite bead, P/N 57-0181.

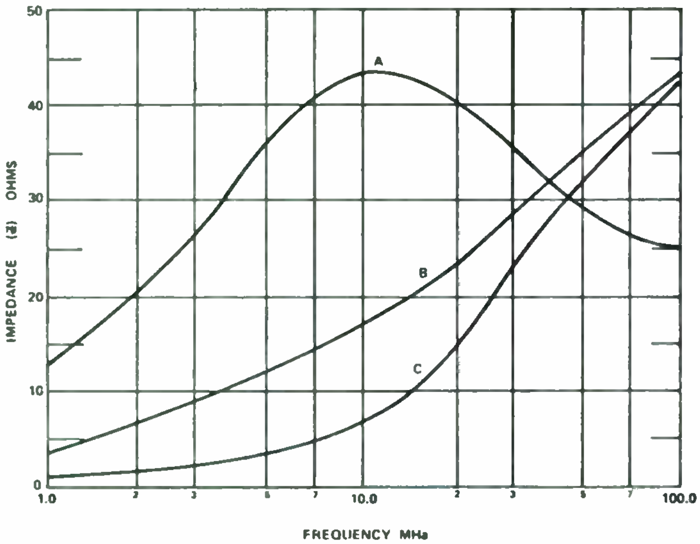

Figure 4. Typical Impedance vs. Frequency curves of three Ceramag® materials using: A — P/N 57-1658 (24); B — P/N 57-0180 (7D); C — P/N 57-9462 (11).

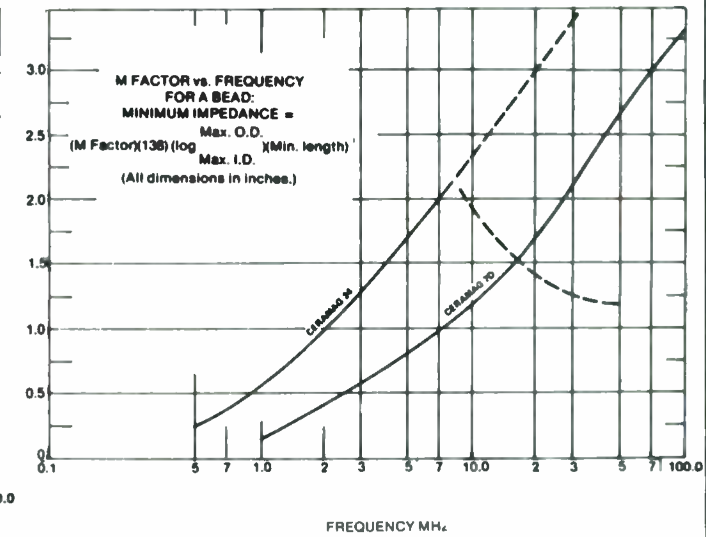

Figure 5. Curves for calculating minimum impedence.

istics, although an M factor is being determined at 100 MHz.

Which material to select?

If the actual noise frequency is known, select the one which provides the most impedance. If the noise is a pulse, the rise time should be considered the frequency.

If an existing bead does not seem to provide the proper impedance,' Figure 5 displays the mathematical factor, M, which is a measure of the minimum guaranteed impedance characteristic of two of Stackpole's materials. These materials have been used in 95 percent of the noise suppression applications up to this time. The formula given in Figure 5 is based on treating the bead, when slipped onto a wire, as a coaxial cavity.

Another factor of interest is that the effect of beads is additive. If one bead provides a certain amount of impedance/suppression, two will provide twice as much and so on. If practical, strings of beads can be used, provided they do not approach one-quarter wavelength of the undesired signal. If at all possible, the O.D./I.D. ratio should between two and three, due to the depth of penetration into the ferrite material by the magnetic field caused by a noise signal surrounding the wire. The frequency at which the

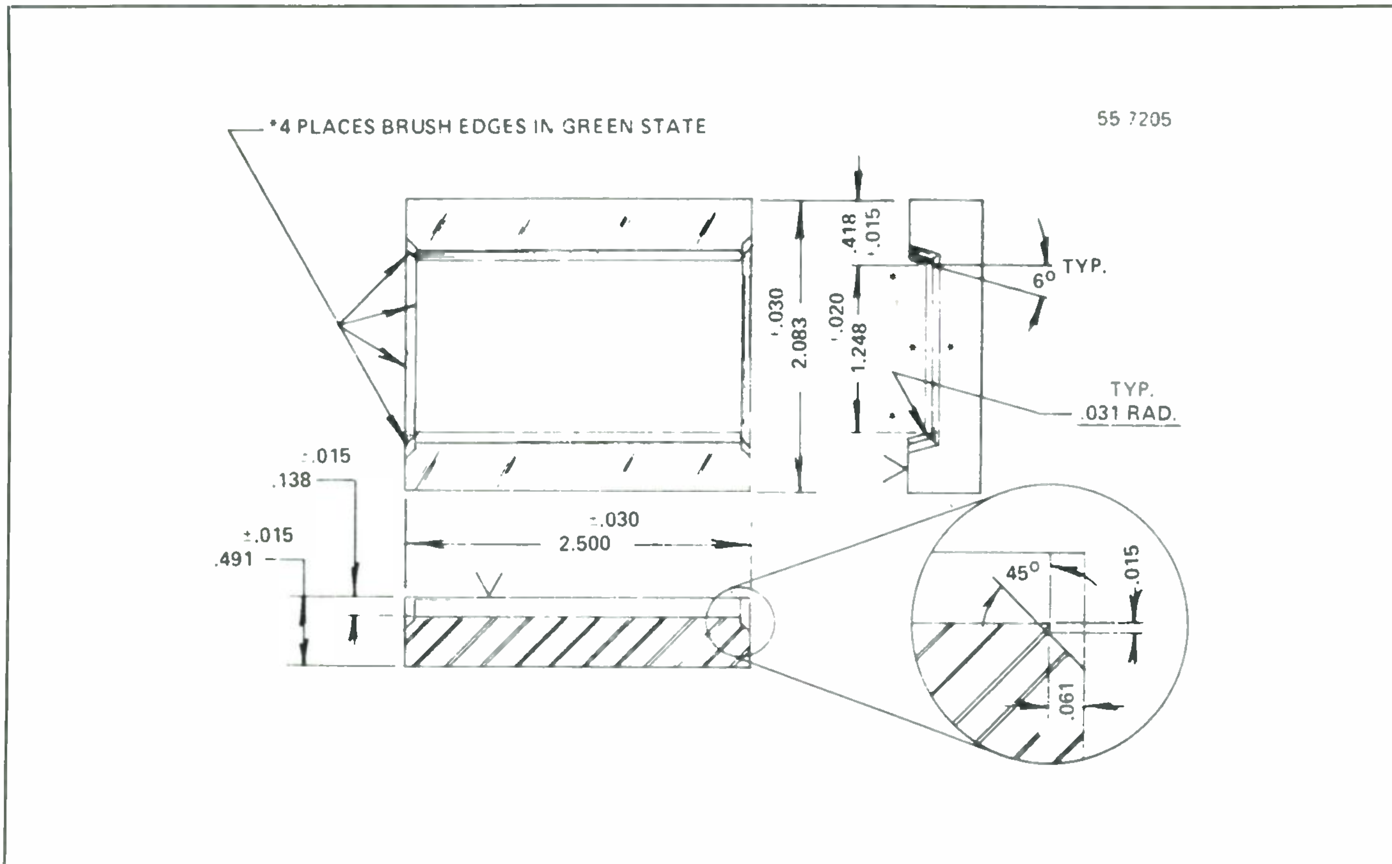

Figure 6. Large bead for flat ribbon cable.

INFO/CARD 34

INFO/CARD 35

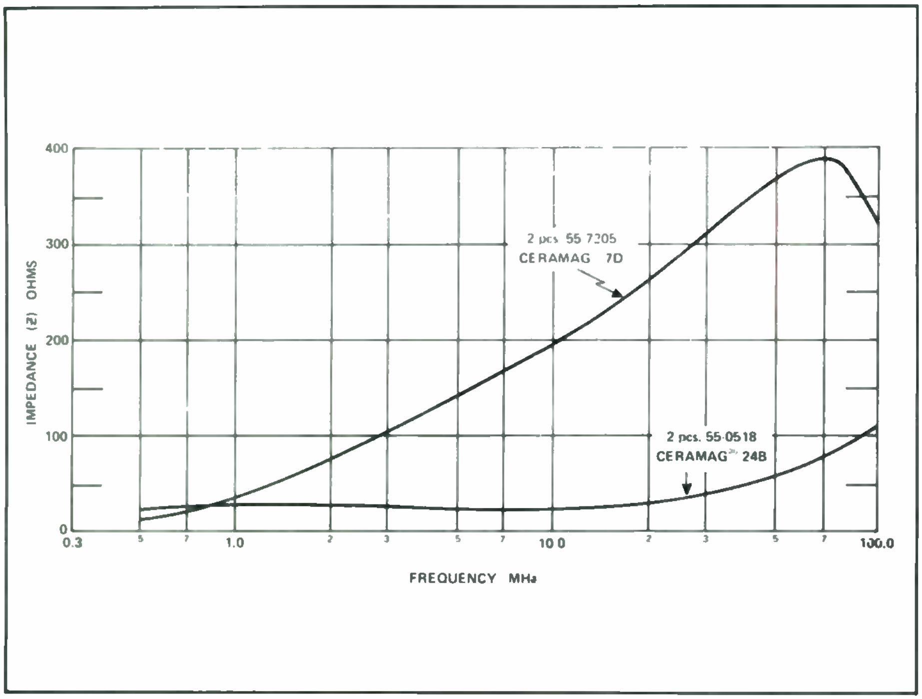

Figure 7. Typical Impedance vs. Frequency curves of two different materials made as per Figure 6, assembled on a 1.187" wide × 0.034" thick × 3.5" long piece of 12 conductor (#23 AWG) cable, shorted together at each end. Measured using HP4815A Vector Impedance Meter.

maximum value of impedance occurs is a material characteristic, while its amplitude is related to the geometry of the core.

If necessary, the wire can be wrapped through the I.D. of the sleeve shape more than once. In such cases, the impedance follows the rule of increasing by the turns-squared factor, each pass through the I.D. being a turn.

Other Forms

The word "beads" has become synonymous with a small cylindrical shape. They can be much larger, as shown in Figure 6.

Developed specifically for flat ribbon cable and available in two different materials, such a configuration can provide suppression even if the cable has to be folded in half to fit within the existing window. The difference in impedance resulting from the use of two grades of material is shown in Figure 7.

This same construction in much shorter versions can also be used as single-turn output chokes on the heavy bus bars of switched-mode power supplies.

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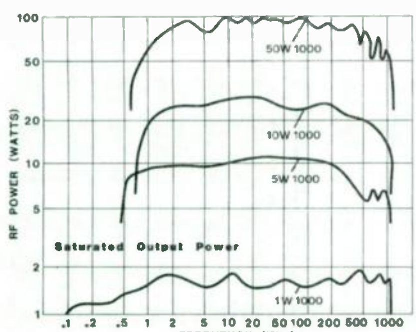

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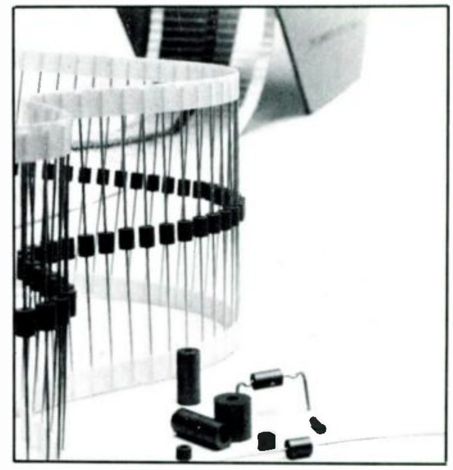

Figure 8.
Beads are available on lead tape and reel for high-volume manufacturing. With the diameter of a ½-watt resistor, they can be automatically inserted on circuit boards.

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Part III

PLL Primer

Part I began the discussion of phase lock loops. Part II considered some design parameters and possible problems. Part III covers further refinements in PLL designs.

By Andrzej B. Przedpelski
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Originally this primer was to consist of Part I (March/April 1983) and Part II (May/June 1983). However, from some of the questions from readers, it became apparent that additional refinements are needed to eliminate the remaining doubts about PLL design. Some of these considerations are necessary only in special cases and will have neglibile effect on majority of PLL designs. However, they may catch the unwary designer off-guard and make his loop oscillate. Since they are very easy to incorporate into the general loop design, it may be safer to build them into the general equations and forget them. Some are mentioned in the literature, but usually as a "fix" rather than an initial design factor. They will be discussed not necessarily in order of importance, since what is important in one design may not be in another. However, where possible, the reason for the specific circuit modification will be described. It should be pointed out that the original third order, type 2, loop (later shown to be actually of a fifth order) will be used as the basis for the discussion. Most of the comments will be also applicable to the better known (but really non-existing) second order loop. The usual integrator configuration for a typical type 2 third order loop is shown in Figs. 6 and 18a. It has been noticed that, when high reference frequencies are used, this configuration does not necessarily provide the expected performance. Changing gain distribution (and readjusting T1) changes performance when it should not. The problem usually is that the integrator operational amplifier cannot handle the very short, high amplitude reference frequency pulses at the input. This can be easily remedied by switching to the configuration of Fig. 18b. The pulses are partially integrated by the passive network RaCa reducing their amplitude and increasing their pulse and rise and fall times. Motorola (Ref. 2) and Gardner (Ref. 15), for instance, consider it mainly as a fix, selecting Ra and Ca to make their time constant neglibile so as not to affect the loop performance. This approach does not provide maximum performance improvement. It is better to use the equivalencies of Fig. 18 (T1, T2 and T3 can be calculated using the program of Table I) and provide maximum passive pulse integration with no effect on the calculated loop performance. This is usually enough to bring the input pulses within the capabilities of the operational amplifier.

Divider Time Delay

While most references ignore time delays in the frequency divider (divide by N), some try to take them into account. Fadrhons (Ref. 16) provides for this delay in his equations, but does not define it. Manassewitsch (Ref. 17) shows the right frequency divider function,

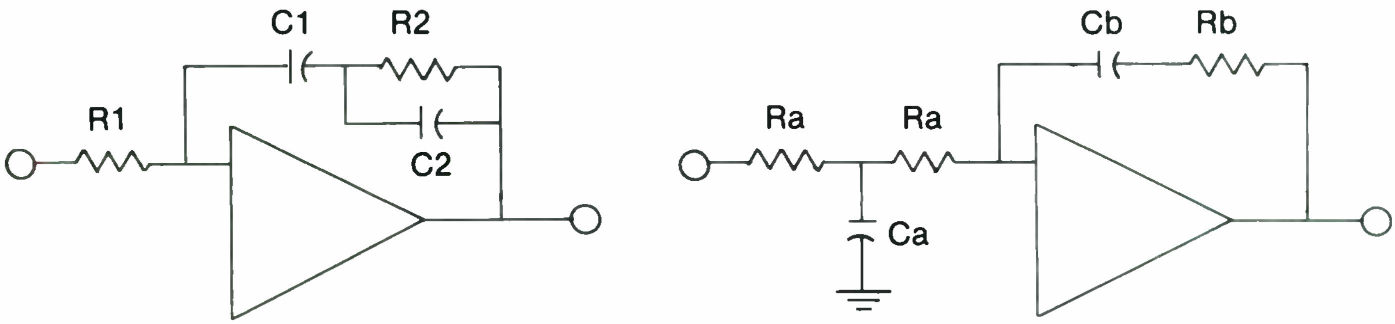
$$G_N(s) = \frac{e^{-ST_N}}{N} \qquad (10)$$

but intimates that T_N is logic switching delay and not inherently due to the division process. He also states that when N is large, the effect can be neglected, while actually, the opposite is generally true. Wetenkamp (Ref. 18) of Watkins-Johnson, on the other hand provides a good analysis, but makes an unnecessary approximation and treats it as a correction factor rather than a design consideration. To summarize, the delay in the divider is not a spurious circuit imperfection, but simply the fact that an output change occurs only after N input changes. In between, the output is in limbo, even though the input may be actually changing phase/frequency. The output won't know it until N pulses are counted. This delay is (N-1) divided by the output frequency. This is equivalent, in the frequency domain, to e^{-ST_N}, where T_N is the time delay. This can be further simplified:

$$e^{-ST_N} = e^{-j\omega T_N} = \cos \omega T_N - j \sin \omega T_N \qquad (11)$$

where $\omega T_N = (2\pi F_o/F_R)(N - 1/N)$

Figure 19 shows the effect of this delay on phase margin (open loop transfer function). It can be seen that if F is chosen to be less than $F_R/100$, the effect is negligible, which is the usual case. Figure 20 shows that the time delay has no effect on open loop transfer function magnitude.



$T1 = R1 \times C1$
$T2 = R2 \times C1$
$T3 = R2 \times C2$

$$F_{(s)} = \frac{1 + s(T2 + T3)}{sT1(1 + sT3)}$$

(a) Conventional

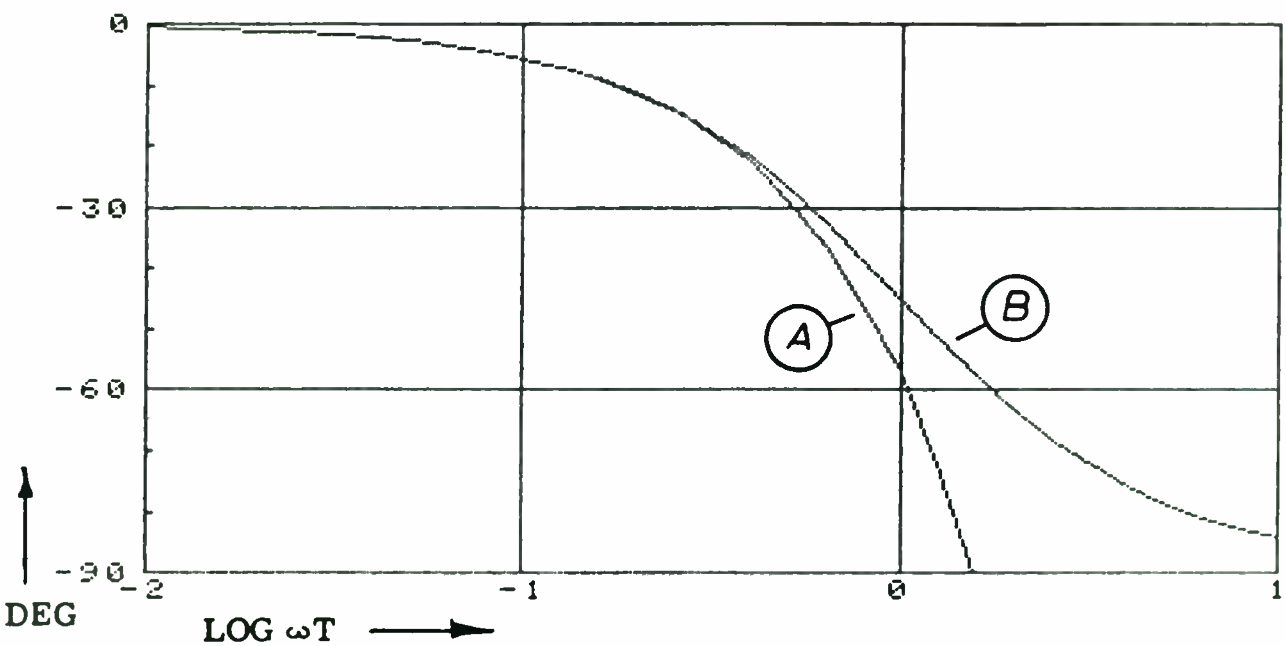
$Tc = T2 + T3$ $Tb = (T\ 1)/2$ $Ta = 2 \times T3$

$Ta = Ra \times Ca$
$Tb = Ra \times Cb$
$Tc = Rb \times Cb$

$$F(s) = \frac{1 + sTc}{2\,s\,Tb(1 + sTa/2)}$$

(b) Modified

Fig. 18. Integrator Configuration.



Ⓐ Exact
Ⓑ W-J approximation

Fig. 19. Open Loop Phase Shift Due To Divider Delay

The open loop transfer function (Eq. 7) can now be modified to include the effect of frequency time delay:

$$G(j\omega)\,H(j\omega) = [\text{Eq. } 7] \times e^{-sT_N} \qquad (12)$$

Note that the closed loop response (Eq. 8) and VTO noise attenuation (Eq. 9) are also modified accordingly.

Effect of VTO Modulation Bandwidth

While Eq. 7, 8 and 9 take into account tne effect of the finite VTO loop bandwidth in calculating the loop performance, Eqs. 4, 5 and 6, used in optimizing the loop, do not. The statement was made that an iterative method could be used, if Eq. 7 indicated that adequate phase margin was not

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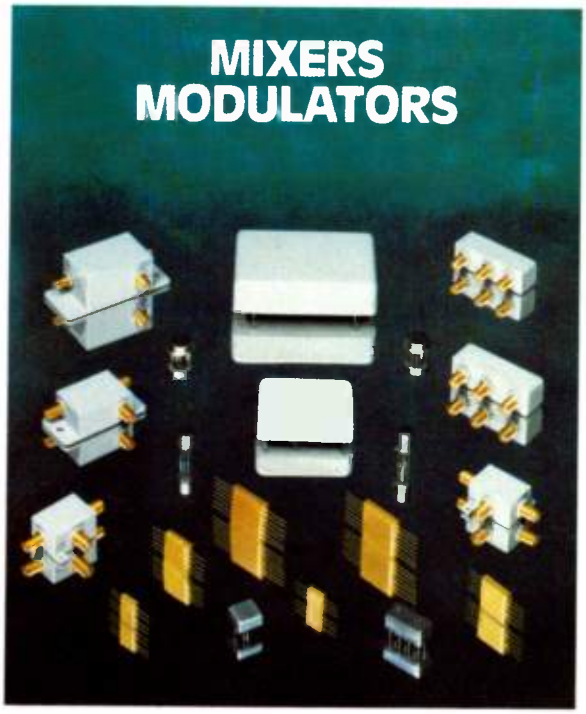

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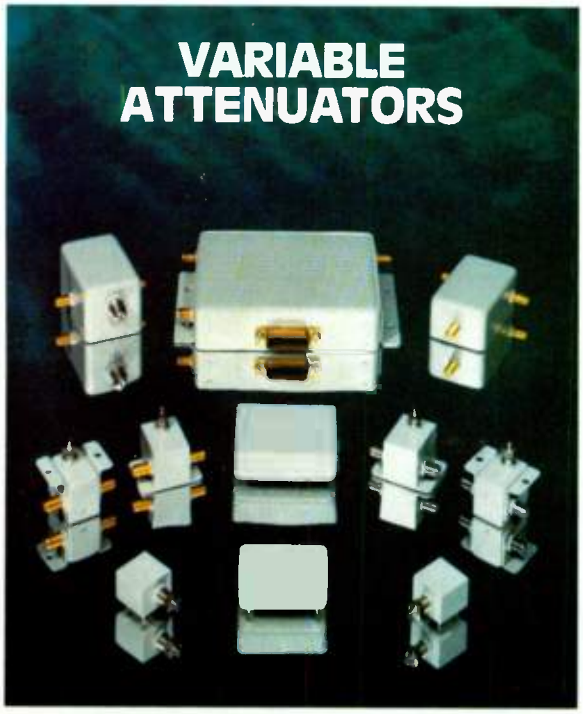

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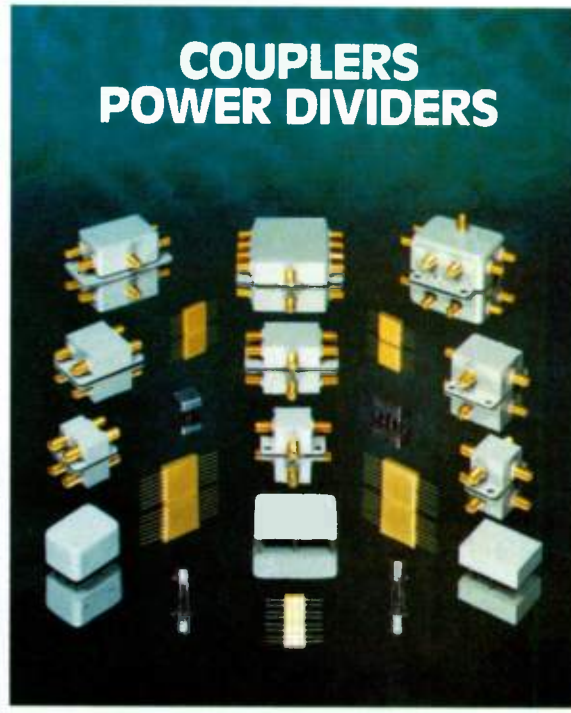

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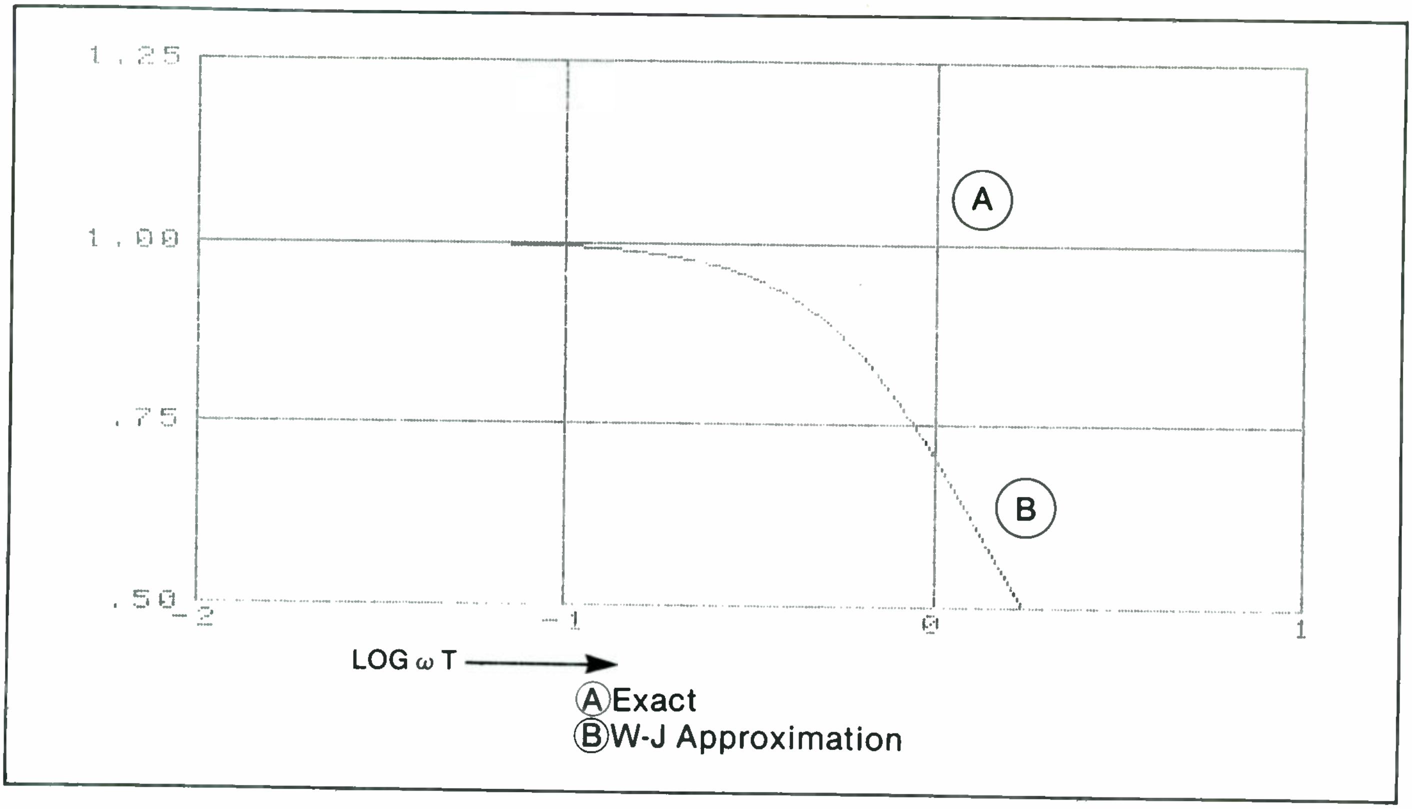

Fig. 20. Open Loop Gain Change Due To Divider Delay.

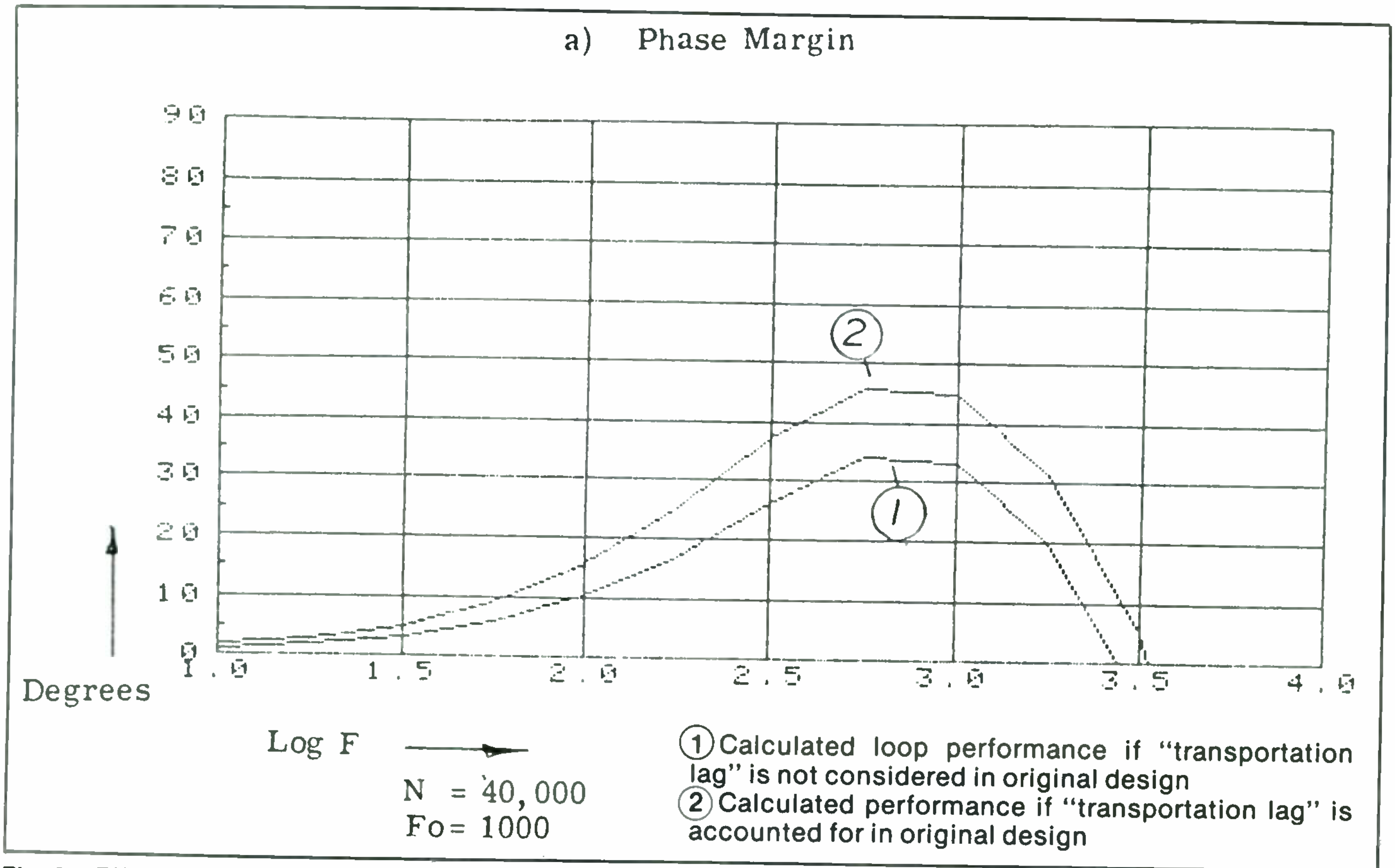

Fig. 21. Effect of "Transportation Lag" on Loop Performance. (Phase Margin)

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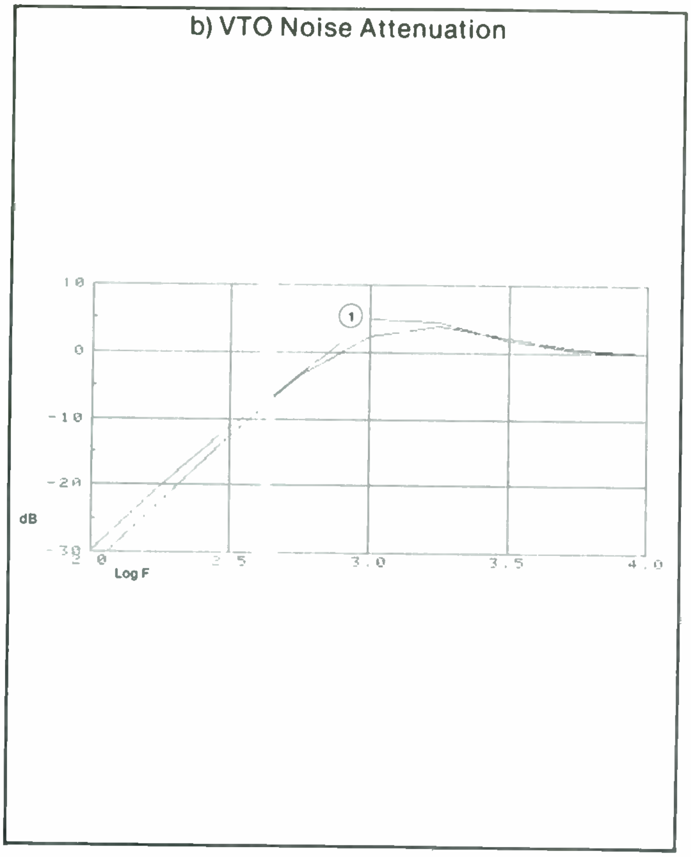

Fig. 21. Effect of "Transportation Lag" on Loop Performance. (V to Noise Attenuation)

available, when the VTO bandwidth was considered. It may be easier to build this compensation into Eqs. 4, 5 and 6. The original method (Ref. 7) was attempted, but differentiating Eq. 7 became somewhat unwieldy. An easier method, used also for the divider delay, was then used. The phase angle used in Eq. 4, 5 and 6 was modified as follows:

$$ø' = ø + \tan^{-1}\frac{F_o}{B_V} + \frac{360\,F_o}{2\pi\ F_R}\left(\frac{N-1}{N}\right) \text{degrees} \qquad (13)$$

where ø is the desired phase margin
This number is then used in Eqs. 4 and 5 for ø. The new equations giving exact solutions for T1, T2 and T3, including the effect of time delay in the divider and the finite VTO bandwidth, become then:

$$T_3 = \frac{\sec ø' - \tan ø'}{\omega_o} \qquad (14)$$

$$T_2 = \frac{2 \tan ø'}{\omega_o} \qquad (15)$$

$$T_1 = \frac{K_p K_v}{N\,\omega_o^2}\sqrt{\frac{1 + [\omega_o(T_2 + T_3)]^2}{1 + (\omega_o T_3)^2}} \times \sqrt{\frac{1}{1 + (F_o/B_v)^2}} \qquad (16)$$

where $ø'$ is defined in Eq. 13
$\omega_o = 2\pi F_o$
B_v = VTO modulation bandwidth

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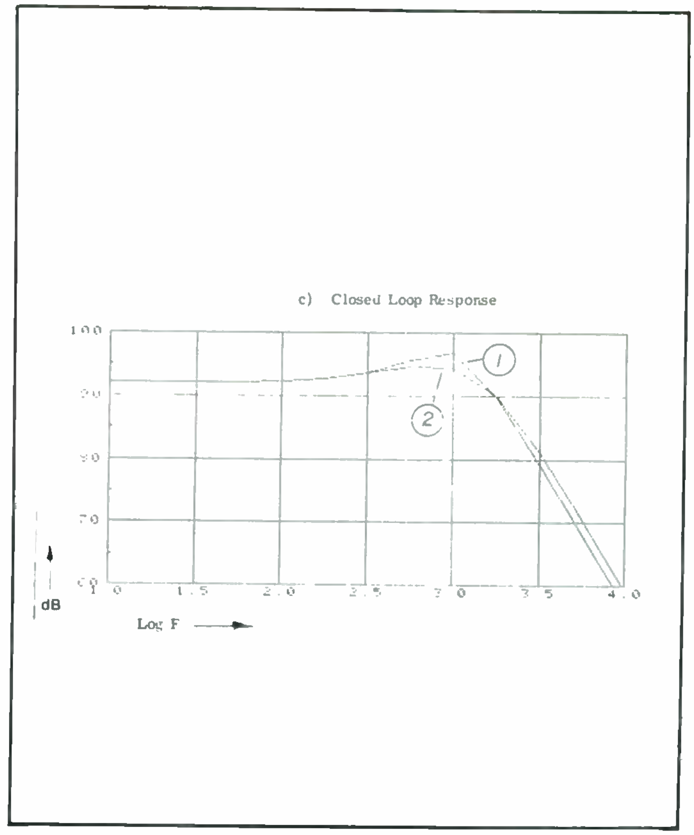

Fig. 21. Effect of "Transportation Lag" on Loop Performance. (Closed loop response)

The loop thus calculated will have the desired phase margin and F_o, but it will not be "optimized", i.e., it will not have the maximum phase margin exactly at F_o, as shown in Fig. 21(a), curve 2. This figure, as well as Fig. 21(b) and (c) shows the consequences of neglecting the effect of the divider time delay on loop performance.

Summary

There seem to be two approaches to PLL design. One, as exemplified by Gardner (Ref. 3) and Motorola (Ref. 2), is to design a classical second order loop and try to reduce all parasitic effects. If necessary, these parasitic effects can be then handled as perturbations. The other, as described in this article, is to account for all effects, even if this increases the order of the loop. These effects are then under the control of the designer and can actually be made use of.

It would be desirable to replace Eqs. 14, 15 and 16 by more rigorously derived ones. Maybe some reader with better mathematical grasp and more patience than the author, can differentiate Eq. 12 and by using the method described in Ref. 7 come up with an exact solution.

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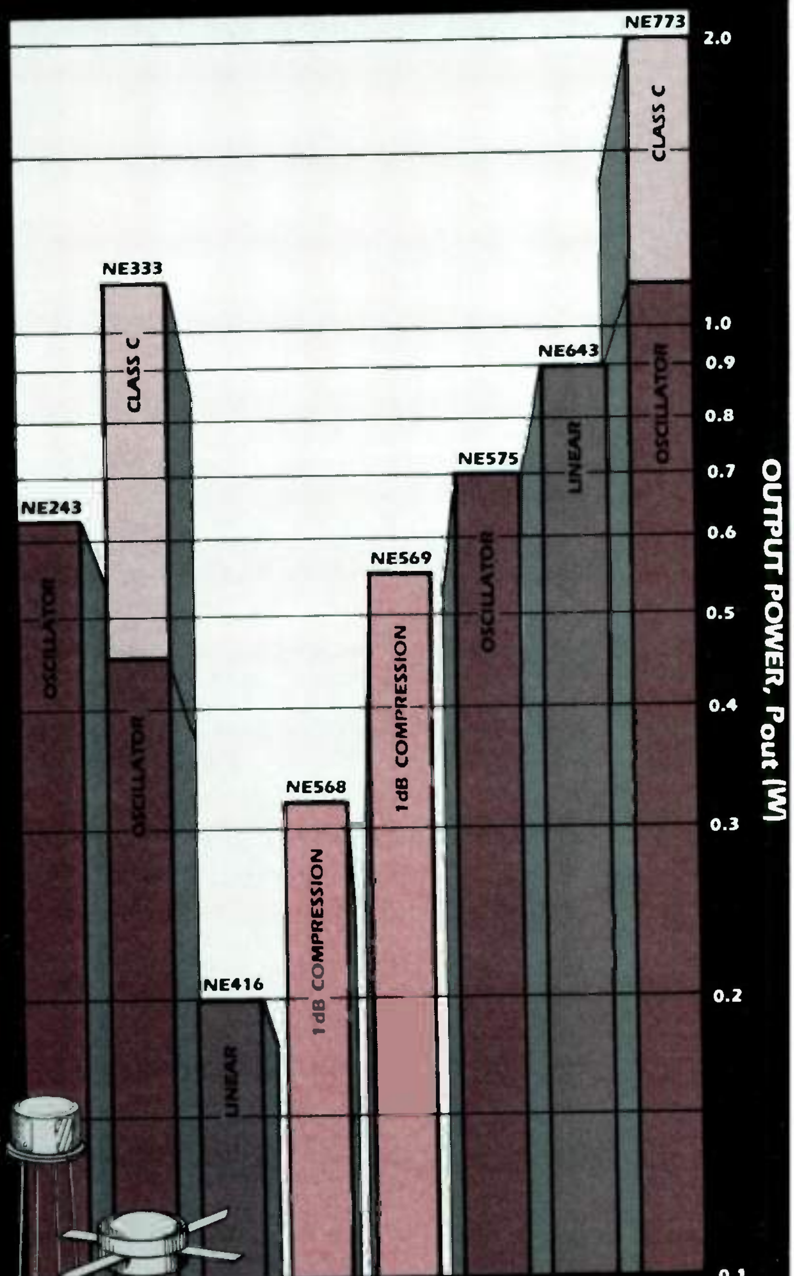

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Electric-Field Measurements Over A Ground Plane

A tutorial approach to the calculation and measurement of electric fields over a ground plane.

By Roger Southwick
IBM General Products Division
Tucson, Arizona

Introduction

The imposition of FCC Regulation Part 15 Subpart J has generated renewed interest in measuring E fields over a conducting ground plane, which is required by this regulation.

The FCC Regulation Part 15 Subpart J specifies limits and methodology for the measurement of digital and other electronic devices. Included is the measurement of undesirable electromagnetic fields that may radiate from such devices.

To better understand and analyze the prescribed measurement methodology, I developed a mathematical model of site attenuation using the conditions imposed by the regulation. I also discuss briefly the use of the model for the application of antenna calibration, fall-off ratios, and site correlation.

Development of Site-Attenuation Model

The model developed is based on modification of the Friis transmission formula [ref. 1]:

$$\frac{W_r}{W_t} = \frac{G_o^2\lambda^2}{(4\pi)^2 d^2} \qquad (1)$$

where:
W_r = Received power
W_t = Transmit power
λ = Wavelength in meters
$G_o^2 = G_1 G_2$, where G_1 and G_2 are the gains of the transmit and receive antennas, respectively
d = Distance in meters

Taking 10 log of Eq. 1:

$$W_r(\mathrm{dBm}) - W_t(\mathrm{dBm}) = 20 \log \lambda - 20 \log d - 20 \log 4\pi + 20 \log G_o \qquad (2)$$

letting $\lambda = 300/f_m$ and f_m = frequency MHz, Eq. 2 becomes:

$$W_r(\mathrm{dBm}) - W_t(\mathrm{dBm}) = 2\,G_o(\mathrm{dB}) - 20 \log d - 20 \log f_m + 27.56 \qquad (3)$$

For the ground-plane condition, a normalized-error* term representing the effect of the reflected wave is calculated using the notation as shown in Figure 1 [ref. 2].

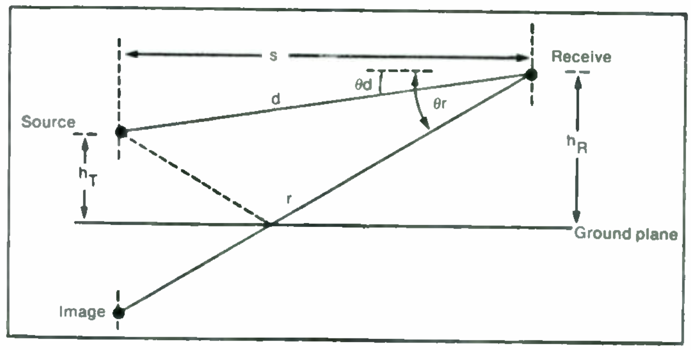

Figure 1. Notation for calculating the effect of a reflected wave.

The vertical, off-axis, correction factor for the biconical antenna is:

$[\cos(\pi L/\lambda \sin\theta) - \cos(\pi L/\lambda)]/[1 - \cos(\pi L/\lambda)\cos\theta] = K\theta$
where L is the overall length of the antenna, in meters.

The vertical, off-axis, correction factor for the log periodic antenna is:

$\cos(\pi/2 \sin\theta)/(\cos\theta) = K\theta$

The equation for the normalized-error term for the case of horizontal polarization is:

$$E_h(\mathrm{dB}) = 20 \log [1 + p^2(d/r)^2 + 2p(d/r)\cos(2\pi(d-r)/\lambda + \varnothing_h)]^{1/2} \qquad (4)$$

and for the case of vertical polarization is:

$$E_v(\mathrm{dB}) = 20 \log [1 + (K^2\theta_r/K^2\theta_d)^2 p^2 (d/r)^2 + 2(K^2\theta_r/K^2\theta_d)\, p\,(d/r)\cos(2\pi(d-r)/\lambda + \varnothing_v)]^{1/2} \qquad (5)$$

The equation for the site attenuation over a ground plane is now:

$$\mathrm{S.A.}(\mathrm{dB}) = W_r(\mathrm{dBm}) - W_t(\mathrm{dBm}) = 2\,G_o(\mathrm{dB}) - 20 \log d - 20 \log f_m + 27.56 + E(\mathrm{dB}) \qquad (6)$$

where E(dB) has been derived for both the horizontal and vertical polarization. The term S.A. (dB) is defined as site attenuation.

Because the FCC regulation specifies dBμv/m as the unit of measure, Eq. 6 is converted to voltage based on a 50 Ω transmission line impedance. Eq. 6 becomes:

* The term normalized error is sometimes referred to as site gain.

$$\text{S.A.(dB)} = E_r(\text{dB}\mu\text{v}) - E_t(\text{dB}\mu\text{v})$$
$$= 2\,G_o(\text{dB}) - 20 \log d - 20 \log f_m$$
$$+ 27.56 + E(\text{dB}) \quad (7)$$

where $E_t(\text{dB}\mu\text{v})$ is the voltage across the terminals of the transmitting antenna, and $E_r(\text{dB}\mu\text{v})$ is the voltage across the terminals of the receiving antenna.

Based on $[(h_t \pm h_r)^2 + s^2]^{1/2}$

$$d = (h_t^2 + h_r^2 - 2\,h_t h_r + s^2)^{1/2}$$
$$r = (h_t^2 + h_r^2 + 2\,h_t h_r + s^2)^{1/2}$$

and the angle of the reflected wave is:

$$\Theta_r = \text{Arctan}\,\frac{h_r + h_t}{s}$$

and the angle of the direct wave is:

$$\Theta_d = \text{Arctan}\,\frac{h_r - h_t}{s}$$

The reflection coefficient is:

$R = pe^{j\emptyset}\, e^{j\emptyset} = 1$ vertical $p = 1$

$e^{j\emptyset} = -1$ horizontal $p = -1$

If the antenna characteristics are specified in terms of antenna factor instead of antenna gain, the following conversion is necessary:

$$G(\text{dB}) = 20 \log f_m - K_a(\text{dB})$$
$$- 29.78 \text{ dB/m} \quad (8)$$

where K_a(dB) is the antenna factor, based on a 50 Ω transmission-line impedance [ref. 3].

Repeatability of E-field measurements over a ground plane is often very poor. The major cause of this problem is the E(dB) term of Eq. 7. This term can, for certain geometries and frequencies, become a large negative value, which is very sensitive to physical change at points of near-total phase cancellation. To avoid this problem, the receive antenna is physically scanned vertically and the maximum received value is measured. To distinguish this scanning process, Eq. 7 is written:

$$\text{S.A.(dB} = E_r(\text{dB}\mu\text{v}) - E_t(\text{dB}\mu\text{v})$$
$$= 2\,G_o(\text{dB}) - 20 \log d - 20 \log f_m$$
$$+ 27.56 + E^{max}(\text{dB}) \quad (9)$$

In the calculation of E^{max}(dB), h_r is incremented over a specified scan range and its maximum value with the corresponding value of d is selected (see Figure 1).

Measurements

The site-attenuation measurements were made using a spectrum analyzer for the receiver and a tracking generator for the signal source. In this combination, the source and receiver track in frequency. The spectrum analyzer must have a "MAX HOLD" function, so that the maximum signal across the frequency and vertical antenna-scan ranges can be measured.

The antennas were positioned in a specified geometry, and the receiver and signal source were connected. With the receiver in "MAX HOLD" and scanning in frequency, the receive antenna was slowly scanned vertically over the specified range. All cable losses were subtracted from the received values by connecting both cables together and measuring $E_t(\text{dB}\mu\text{v})$ at the output.

A three-meter antenna separation, with a transmitter-antenna height of one meter, and a vertical receive-antenna scan of one to two meters were used.

Measurements were made at the U.S. Army Electronic Proving Ground, Ft. Huachuca, Arizona, antenna range. This range was used as a standard of reference. It has an extensive ground plane and an instrument room under the ground plane. A pair of Watkins Johnson, log-periodic antennas, model AR12-20, was used in the 200-1000 MHz frequency range, and a pair of biconical antennas was used in the 30-200 MHz range. Antenna-factor data used to calculate site attenuation was based on measured data provided by R. German, IBM Boulder [ref. 4].

The absolute value of the mean plus twice the standard deviation of the error was used as a criterion. The error was the difference between the measured and the calculated site-attenuation data. The mean and standard deviations were calculated across the frequency range. The criterion is used in Figure 2, under the heading "Error (dB)."

The results of comparing the measured and calculated site-attenuation data are shown in Figure 2. In every case the criterion Error(dB) was less than 3 dB, which is adequate validation of the approximation of Eq. 9 to measured site-attenuation data taken at a standard site.

Frequency Range (MHz)	Polarization	Mean Error (dB)	Std. Dev. (dB)	Error (dB)
200-1000	Vertical	0.228	1.236	2.701
	Horizontal	0.636	0.833	2.303
30-200	Vertical	0.290	1.220	2.644
	Horizontal	0.126	1.259	2.731

Figure 2. Comparison of Measured and Calculated Site-Attenuation Data.

Figures 3 to 6 show the data plots corresponding to Figure 2.

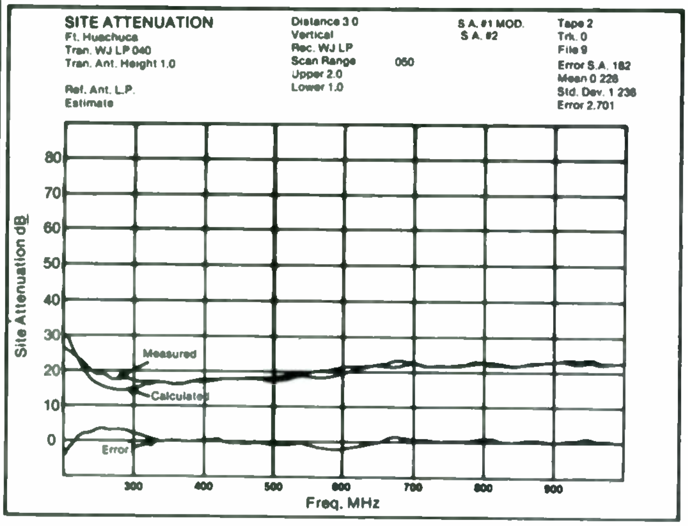

Figure 3. Data Plot for 200-1000 MHz, Vertical Polarization

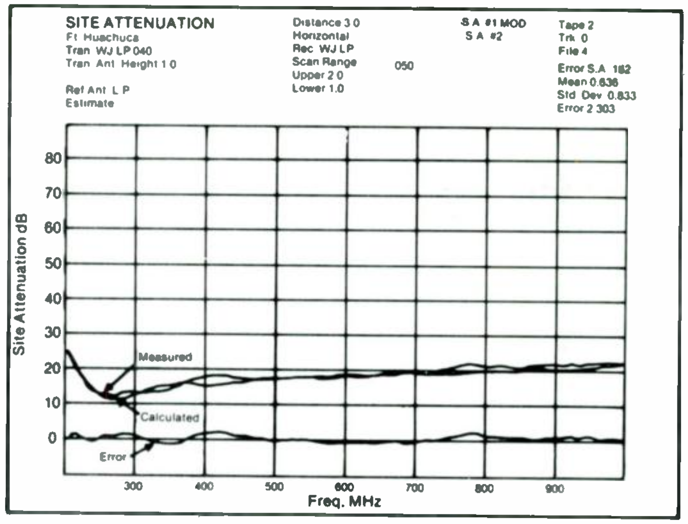

Figure 4. Data Plot for 200-1000 MHz, Horizontal Polarization

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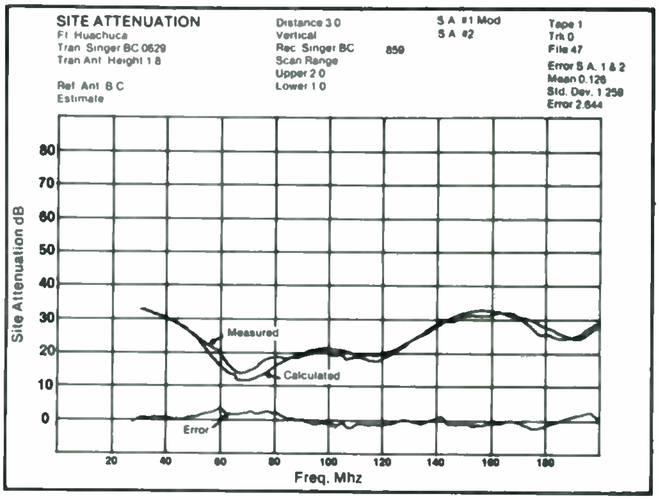


Figure 5. Data Plot for 30-200 MHz, Vertical Polarization.

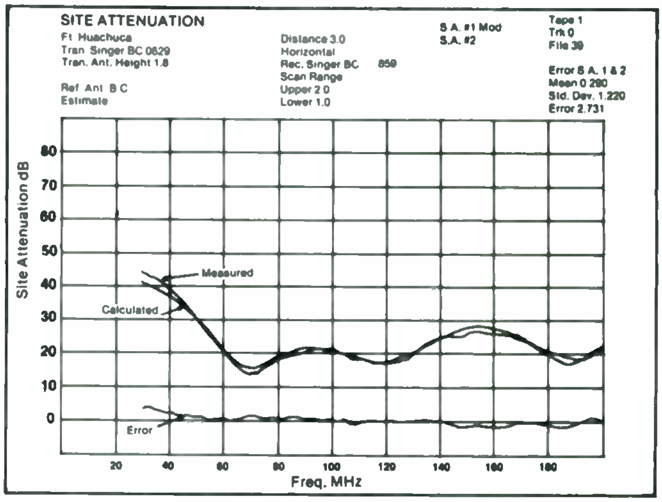


Figure 6. Data Plot for 30-200 MHz, Horizontal Polarization.

Software was written for a desk-top computer to measure and calculate site-attenuation data in accordance with Eq. 9. This software also includes extensive plotting routines that permit selective plotting of both measured and calculated site-attenuation data and their difference, antenna factor and gain, plus the E(dB) term. Provision for recording the measured data was also provided. The data plots in Figures 3 to 6 were generated by means of this software.

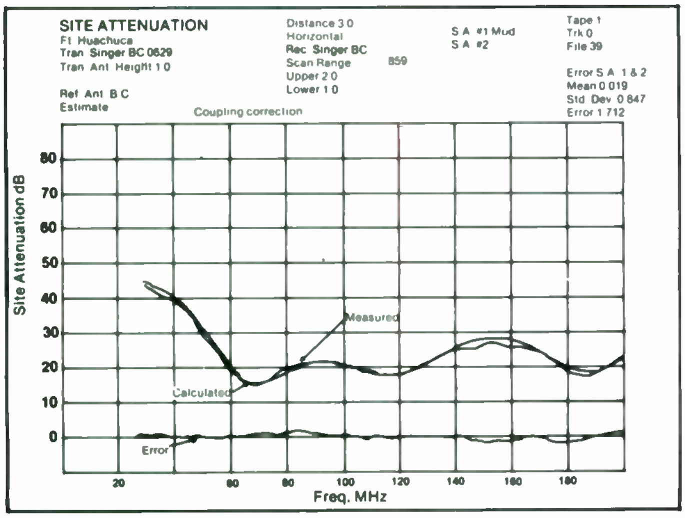


Figure 7. Data Plot for 30-200 MHz, Horizontal Polarization with Coupling Factor.

All data is plotted with the sign of Eq. 9 reversed so that the data will agree with previously established conventions.

Antenna Coupling And Calibration

When site-attenuation measurements are made over a ground plane, site coupling of the antennas to each other and their images occurs [ref. 5]. The effect of site coupling is a change in the antenna impedance, which results in a change in antenna gain. This effect is not included in Eq. 9. The site-coupling effect is related in general in a nonlinear way to the separation of the antennas and their images. The most severe condition is the image coupling for horizontal polarization at lower frequencies. This effect has been documented by actual measurement [ref. 8] and also by the author.

It would appear that, since the receive antenna must be scanned vertically and the coupling to its image is a function of distance, the determination of a site coupling's connection factor would be impossible. A computer simulation using Eq. 9 for horizontal polarization, however, shows that in the frequency range of the most severe coupling, the maximum received signal will always occur at the maximum receive-antenna height and, therefore, it is only necessary that a site-coupling-antenna correction factor be determined for the transmitter antenna. When the site-coupling-antenna correction factor is included in Eq. 9 for horizontal polarization, a more accurate site-attenuation measurement will result.

An optional site-coupling factor for horizontal polarization based on measured data was developed. The factor is a straight-line approximation of the error with its slope inversely proportional to the height of the transmitter antenna. This factor was used to plot the data shown in Figure 7, which corresponds to Figure 6 where the coupling factor was not used.

Antennas can be calibrated over a ground plane by solving Eq. 9 for G_o and using measured values of site attenuation [refs. 6 and 7]. It is necessary to provide adequate antenna separation to minimize near-field effects and to accept the site-coupling error in the resultant antenna factor.

I wrote a subroutine that permitted the error difference between measured and calculated site-attenuation data to be analyzed for both vertical and horizontal polarization. This subroutine corrects the antenna factor so that the error of both polarizations will either remain unchanged or decrease. The corrected antenna factor that results is then stored for future use. The results of optimizing the antenna factors and applying the coupling correction factor for horizontal polarization are shown in Figure 8.

Frequency Range (MHz)	Polarization	Mean Error (dB)	Std. Dev. (dB)	Error (dB)
200-1000	Vertical	0.017	1.009	2.035
	Horizontal	0.425	0.655	1.735
30-200	Vertical	0.111	0.857	1.825
	Horizontal	0.004	0.454	0.912

Figure 8. Comparison of Measured and Calculated Site-Attenuation Data with Optimized Antenna Factor and Coupling Correction Factor. Horizontal polarization 30-200 MHz.

The error term has now been reduced to approximately 2 dB or less, which is further verification of the correctness of the model.

Application to Fall-Off Ratio

Another application of Eq. 9 is the analysis of sets of measured site-attenuation data. This was accomplished by additional software, which compares two sets of measured site-attenuation data. Of interest is the comparison of site-attenuation data taken at different distances, because the FCC regulations requires that data be extrapolated to thirty meters.

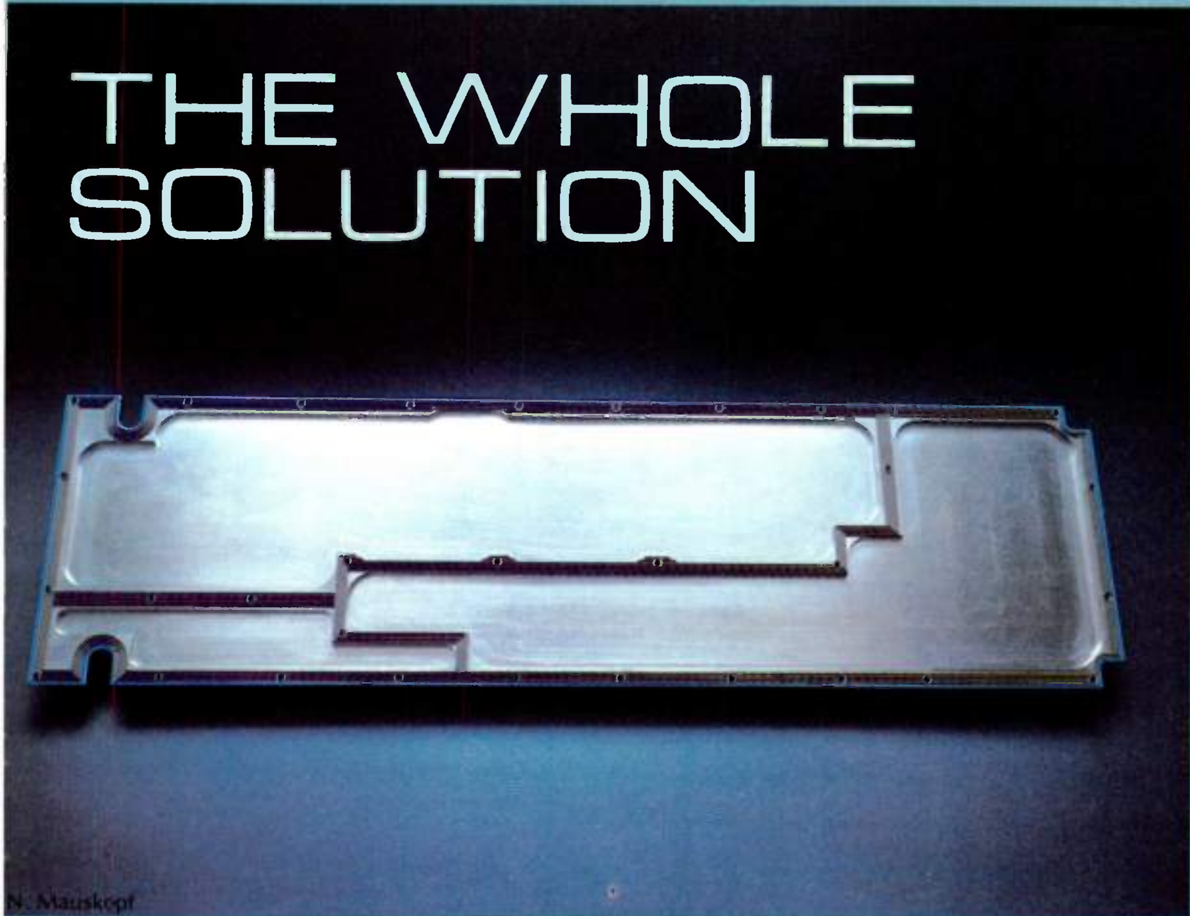
When site-attenuation measurements are made at different distances, the difference, Δ S.A. (dB), can be expressed as:

$$\Delta \text{ S.A. (dB)} = \text{S.A. (dB)}_1 - \text{S.A. (dB)}_2$$
$$= -20 \log d_1/d_2 + E(dB)_1 - E(dB)_2 \qquad (10)$$

The right side of Eq. 10 must equal 20 log s_1/s_2 so that the latter fall-off ratio can be applied. The terms d_1 and d_2 are determined by the value of h_t when the maximum received signals are measured. The E(dB) terms contain the site's geometry dimensions, including h_t, which is unknown for an actual equipment under test (see Eqs. 3 and 4). The E(dB) terms will only be equal when the received signal is at a theoretical maximum, and, in such cases, the receive antenna will most probably not equal the source height, so that $d \neq s$. Extensive measurements [ref. 9] demonstrate the futility of attempting to extrapolate E-field data using the 20 log s_1/s_2 fall-off equation.

Site Correlation

Measured site-attenuation data taken at different sites can also be used to evaluate sites. Data taken at an ideal



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open-field site can be used to evaluate a radio-frequency, semianechoic-chamber site [ref. 10]. The comparison of measured and calculated site-attenuation data provides one measure of error. The error caused only by site differences is obtained by comparing the measured site-attenuation data of a reference site with that of an alternate site.

Conclusion

The use of the E-field model of Eq. 9 with an automated measurement system is a useful tool in the analysis of site attenuation. One advantage of this method is that sweep-frequency, site-attenuation and antenna-gain measurements can be easily made without either a standard antenna or a standard field.

A second advantage is that a computer simulation using Eq. 9 will provide a convenient method to analyze the effects of site geometry on measurement errors.

Acknowledgments

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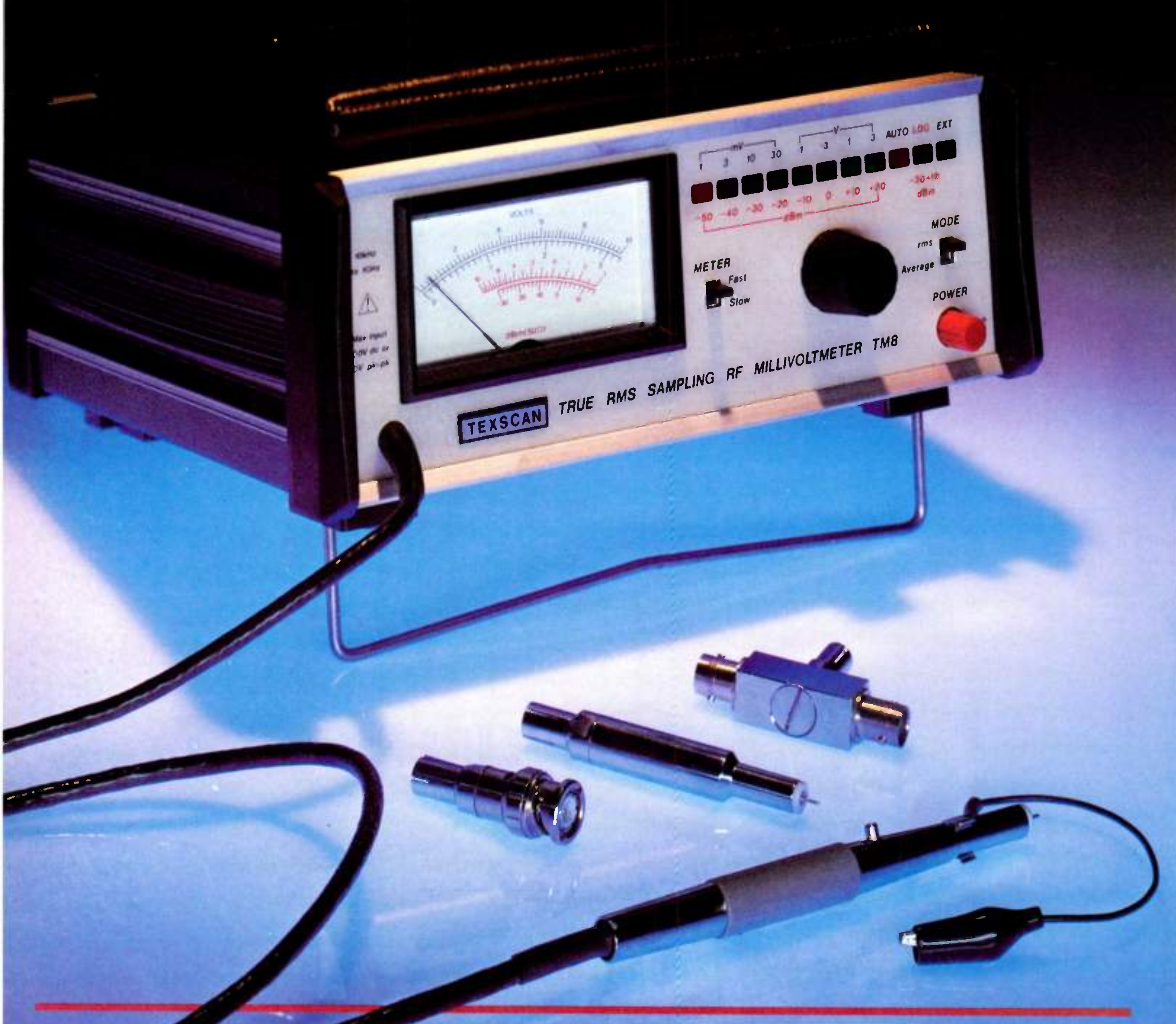

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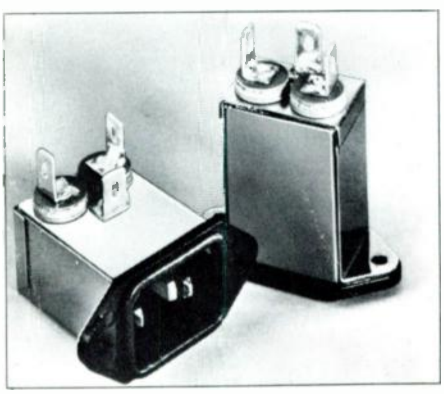

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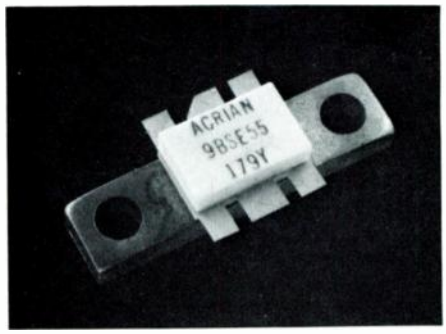

the industry. A single 9BSE55 can now be used in place of a pair of 30 watt devices to achieve the same power output. The 9BSE family is currently in volume production and ready for immediate shipment. Pricing information is available upon request. **Acrian, Inc., Cupertino, CA 95014 or INFO/CARD #167.**

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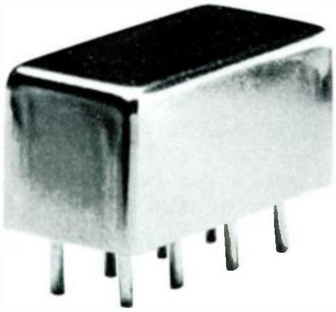

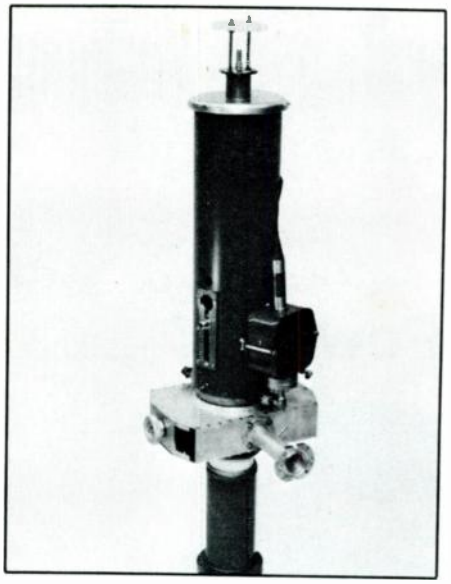

60 percent. Power gain is much higher than that of conventional tetrodes, and the fact that the tube operates with an HV power supply having the positive end grounded makes it particularly suitable for use with a klystron visual amplifier, which uses a similar positive-ground configuration. The Klystrode can operate at the same beam voltage as the associated visual klystron, so that a common beam power supply may be used, subsequently reducing cost. Varian's Klystrode appears to be well suited for visual UHF-TV service. Because the beam input power varies with video modulation level, the klystrode is very energy-efficient. It is presently offered as an aural amplifier. In either service, the high power gain, efficiency, and expected long life are certain to attract the attention of broadcast designers. **Varian Eimac, San Carlos, CA 94070 or INFO/CARD #171.**

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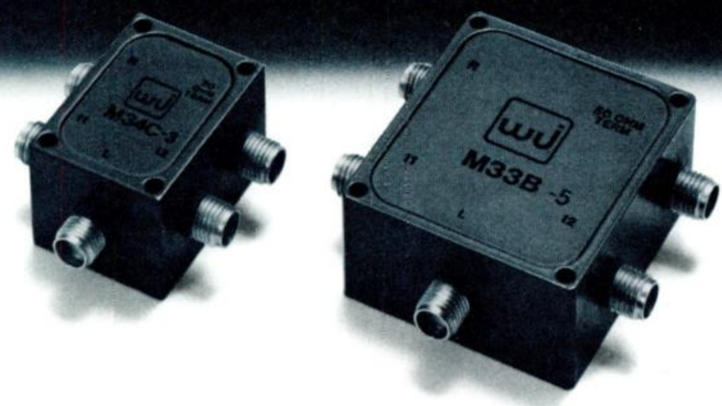

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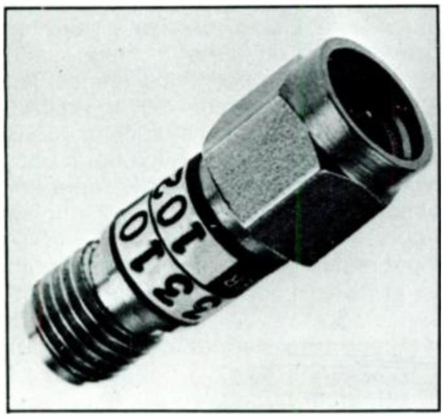

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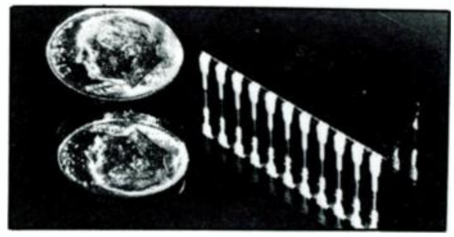
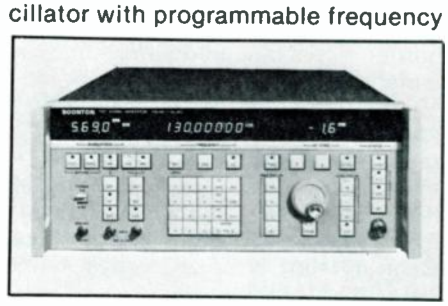


build a practical receiver. These include two L/C filters, one ceramic filter, capacitors, potentiometers, and a quartz crystal. The S 469 is specified for operation between −30° to +60°C. Other typical functional specifications include; HF prestage — voltage gain from 10 to 50 MHz of 36 dB, input impedance of 10 kohns/3pf, and a noise figure of 6 dB; IF limiter amplifier — input resistance of 2 kohms, minimum bandwidth of 500 kHz, AF output voltage of 60 mV, AF bandwidth of 35 kHz, and total harmonic distortion of 1%; AF amplifier #2 — voltage gain of 37 dB, input impedance of 10 kohms, signal to noise ratio of 40 dB, and typical harmonic distorition of 1%, muting depth range of 10 to 40 dB (maximum). Deliveries are stock to 6 weeks. In 100 lots, prices are $4.75 each. **Siemens, Iselin, N.J. 08830 or INFO/CARD #107.**

Programmable Signal Generator For Cellular Receivers

Boonton Model 1021 programmable r.f. signal generator, with a frequency range to 1.08 GHz, covers the requirements for high speed testing of cellular receivers. Switching time between channels is 50 ms. Output levels to +16 dBm, SSB noise of less than −113 dBc, residual f.m. below 12 Hz at 900 MHz (300 Hz to 3 kHz B.W.), and f.m. distortion of 0.05% are provided. The low distortion, internal, modulation oscillator can be used as an audio oscillator with programmable frequency



Now, HP packs more value into its economical RF spectrum analyzers with digital display and rugged portability.

Measuring RF signals? HP's economy RF spectrum analyzers now give you even more value for your money. We've added digital display, versatile data I/O and rugged portability to the excellent performance and operating simplicity of these popular spectrum analyzers. The new HP 853A digital display spectrum analyzer mainframe with the HP 8557A (10 kHz-350 MHz) or 8558B (100 kHz-1500 MHz) spectrum analyzer plug-in combines accurate performance and simple three-knob operation to meet a wide range of RF measurement applications.

Digital display adds clarity and convenience.

The HP 853A provides two independent 480-by 800-point resolution traces for flexibility in storing trace data and monitoring signal changes. You get a crisp, flicker-free, easy-to-read display. And the microprocessor-managed display system includes maximum hold for storing peak signal values, digital averaging over successive sweeps to reduce noise effects, and trace normalization to help detect signal changes in a crowded spectrum.

Rugged portability lets you take it almost anywhere.

The 853A is designed for transportable use in rigorous environments. Weighing less than 48 lbs., the analyzers can be taken outdoors for field test use, or easily moved around lab and production sites. The 853A comes with a tilt-bail handle, rubber bumpers, and a drip-proof front panel cover. What's more, the HP 853A with plug-in installed is type-tested to meet shock, vibration and driptest levels specified under MIL-T-28800 C Type III, Class 3, Style C performance tests.

A choice of RF plug-ins covers your applications.

Notable performance specifications of the 8557A and 8558B plug-ins include 70 dB dynamic measurement range, 1 kHz to 3 MHz variable resolution, full range frequency response of ±1 dB and 70 dB IF substitution amplitude accuracy of ± 1 dB. Basic measurements require only a three-knob sequence: TUNE to a signal and read frequency from the LED display, narrow the FREQUENCY SPAN for closer analysis and adjust REFERENCE LEVEL to read amplitude directly. Resolution and sweep speed are automatically adjusted.

U.S.A. list prices for the plug-ins are: HP 8557A (350 MHz) $6,520; HP 8558B (1500 MHz) $7,925. The new 853A display mainframe is listed at $5,550.*

For a permanent record or comparison of many displays, the 853A push-button controls can directly operate an HP-IB plotter for hard copy data recording. Or connect the 853A to a computer (like the HP-85) via HP-IB and get trace I/O as well as operator prompts on the CRT.

DESIGNED FOR HP-IB SYSTEMS

To add portability and digital display convenience to your RF spectrum analysis applications, call your nearest HP sales office and ask for "Instruments." For the same simplicity and performance in microwave applications, ask about HP's economical microwave plug-in for the 853A. For more information about these instruments, write Hewlett-Packard, 1820 Embarcadero Road, Palo Alto, CA 94303.

*U.S.A. list prices only

5301201

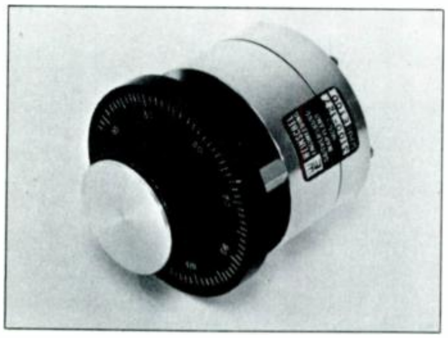
HEWLETT PACKARD

INFO/CARD 74

and level. Control may be manual, automatic via GPIB, or preset to recall up to 250 complete panel setups from an integral non-volatile memory. Price: Model 1021, $16,950, availability: 4 - 6 weeks. **Boonton Electronics Corp. Randolph, NJ 07869, or INFO/CARD #105.**

Smaller Variable Attenuator

OEM designers will have greater flexibility designing compact instrumentation with Weinschel's new model 3100 attenuator. This miniature attenuator weighs only 8.75 ounces and is the smallest available precision attenuator with an incremental attenuation range of 127 dB. Its depth behind the panel is less than 2.5 inches.



Continously variable from 6 dB to 133 dB, the direct reading dial is calibrated in 2 dB increments. Price: $475, delivery: Stock to 90 days. **Weinschel Engineering, Gaithersburg, Maryland 20877, INFO/CARD #106.**

Coaxial Switches

RLC Electronics has introduced a line of Terminated Multi-Position Coaxial Switches, which provide proven reliability, long life and outstanding electrical performance, where termination of output switch lines is required. Electrical characteristics feature extremely low insertion loss and VSWR over the entire DC-18 GHz range, while maintaining high isolation. The switch is available in remote, 28 Vdc or 115 Vac operation, with or without indicator terminals, failsafe or latching. Standard RF power rating is 2 watts CW, limited by the termination. **RLC Electronics, Inc., Mt. Kisco, New York 10549, or INFO/CARD #104.**

Diode Chips

New England Microwave Corporation, is now offering their 14000 Series PIN and NIP chips for high speed switces and attenuators operating over the 1 MHz to 18 GHz frequency range. Suitable for MIL-STD 883 components, these diodes feature a broad range of junction capacitance down to .01 pf and permit switching speeds down to 1 nanosecond. Both single and multiple mesa configurations will bond easily with either ribbon or wire. **New England Microwave Corporation, Hudson, New Hampshire 03051, or INFO/CARD #102.**

90 To 120 MHz Multiple Crystal Oscillator Module

CTI has developed a family of 2 to 24 channel switchable multiple-crystal oscillators for applications requiring high spectrual purity, extremely low AM and phase noise characteristics and ultra good frequency stability. The basic module consists of multiple crystal oscillators (expandable to 24 oscillators) that are BCD selectable. Each oscillator has ultra low phase noise of 120 dBc/Hz at 100 Hz, 135 dbc/Hz at 1 KHz and 149 dbc/Hz at 10 KHz. Each crystal oscillator is automatically

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IE ☐ Instrumentation and test equipment mfr
CE ☐ Consumer electronic mfr
ME ☐ Marine Communication equipment mfrs
GV ☐ Government Military, Aerospace, ECM, EW radar mfrs
NE ☐ Navigation Equipment mfrs
CM ☐ Components mfrs
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phase stable 5 or 10 MHz internal reference or an external 5 or 10 MHz frequency standard. Each selectable output frequency has the ultra low noise characteristics of the 100 MHz range crystal oscillator and the excellent long term frequency stability of either the internal or external reference. The phase noise of the output is not effected by that of the reference. **Communications Techniques, Inc., East Hanover, NJ 07936, or please circle INFO/CARD #103.**

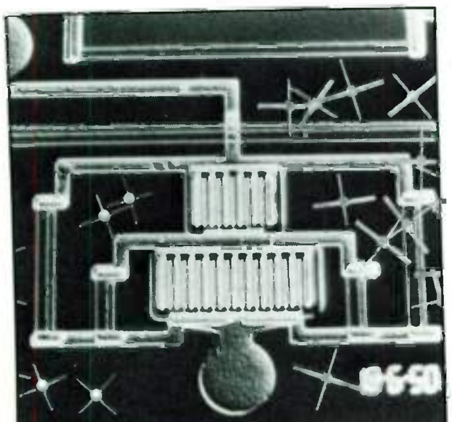
Low Attenuation Coax Cables

Belden Electronic Wire and Cable has available three 50 ohm low attenuation, flexible coax cables (Belden 9913, 9914, 9915) for cellular radio, satellite communications, microwave and other two-way communications.

The cables are designed as flexible alternatives to semi-rigid cable to allow for ease of installation while maintaining similar electrical parameters. Due to construction techniques, these cables are lower in attenuation than other flexible coax cables of the same size. The cables will fit standard connectors. **Belden, Geneva, IL 60134. INFO/CARD #101.**

MMIC Modular Amplifiers

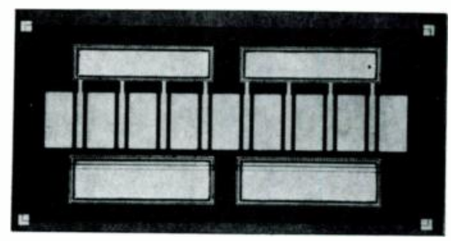
Avantek Inc., Santa Clara, CA, has released the first three of a series of silicon, microwave monolithic modu-



lar integrated circuit amplifiers (MMICs). Available in either military qualifiable 70 mil or economical micro-X transistor packages, MODAMP™ cascadable amplifiers offer up to 19 dB of gain, 5 or 6 dB noise figure, 1 dB power output levels of + 1.5 to + 10 dBm and third order intercept points of + 15 to + 23 dBm (all typical). Each model is usable up to 3 GHz with VSWR of less than 2.0:1 over its entire operationg range. Negative feedback results in good thermal stability from − 55 to + 125°C and group delays on the order of 150 picosec. Internal biasing circuitry allows operation from a single bias source and they are electrically rugged and forgiving of reverse bias, overvoltage, RF overdrive and electrostatic discharge. The high gain and good performance of these MODAMP silicon MMICs is achieved through the use of both series and shunt feedback making them compact, low cost alternatives to hybrid amplifiers. They should find wide acceptance in military and commercial systems as both narrow and broadband RF and IF amplifiers as their various grounding and biasing options make them easily adaptable to most circuit requirements. Available from stock, domestic U.S. prices range from $11.50 and up in micro-X packages and $16.25 and up in the 70 mil package. **Avantek Semiconductor Sales, Santa Clara, CA 95051, or INFO/CARD #100.**

Ku-band GaAsFETs

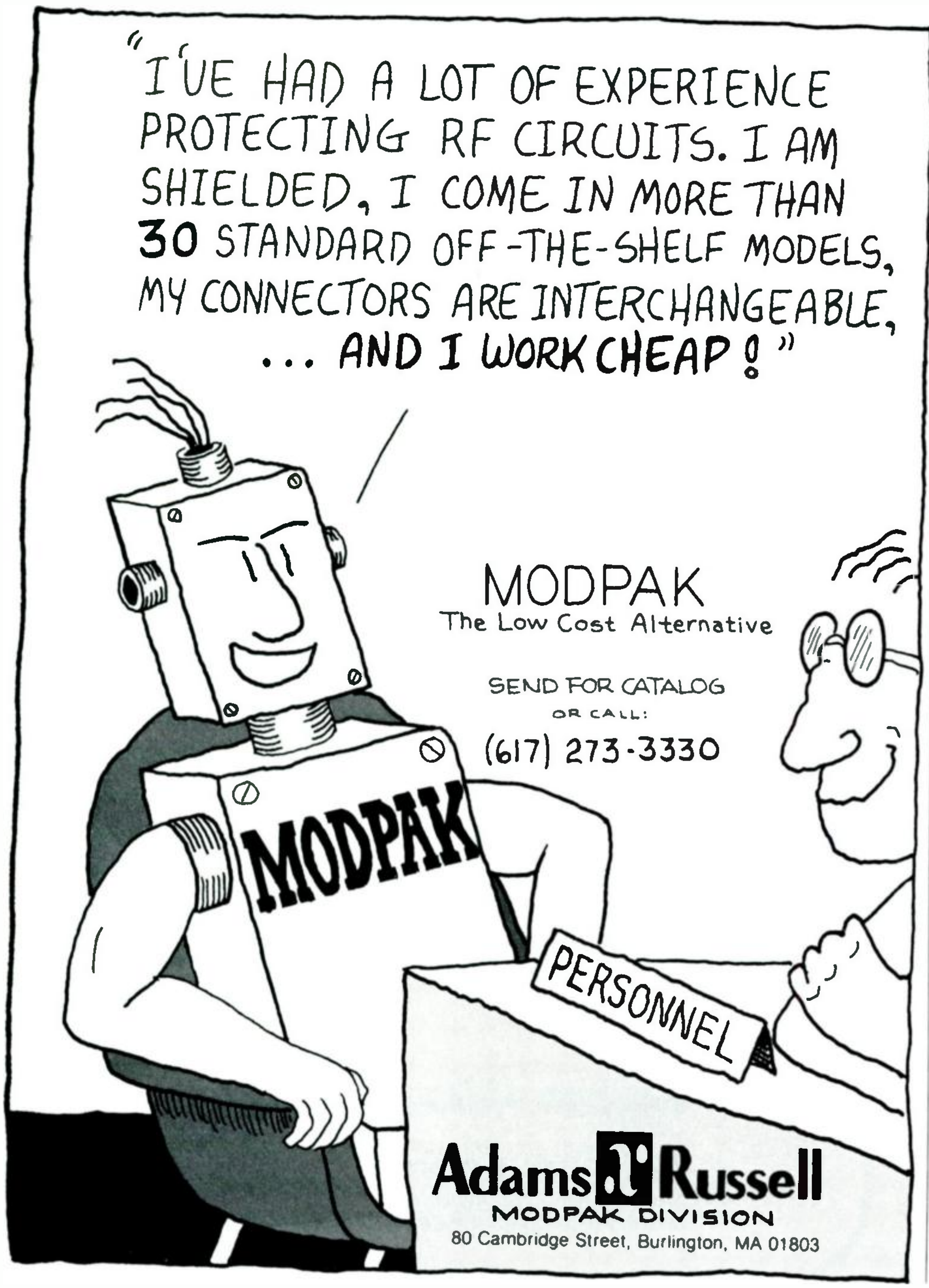
Two new power GaAs field-effect transistors for operation at Ku-band and featuring low inductance via-hole source connections are now available from Raytheon's Special Microwave Devices Operation. The RPK9027 and RPK9030 deliver 0.5 watt and 1.0 watt, respectively, at 18 GHz with 5 dB associated gain. Both devices are obtained from the same chip by utilizing either one or two cells. Each cell consists of eight aluminum gate fingers measuring 100 microns wide by 0.7 micron long. Gate and drain



pad metallization is titanium-platinum-gold. Source bond wires are not required on RPK9027 or RPK9030 chips because of the use of via-hole source connections. At the same time, via-hole source connections result in highly stable operation by virtually eliminating spurious oscillations. Both devices are available either as chips, mounted on Type 100 miniature carriers, or on customer-specified carriers. Small quantity pricing of RPK9027 chips is $100.00 each, while RPK9030 chips are $200.00 each. Chips are available from stock. **Raytheon Company, Special Microwave Devices Operation, Northborough, MA 01532, or INFO/CARD #174.**

GaAs Low Noise FET's

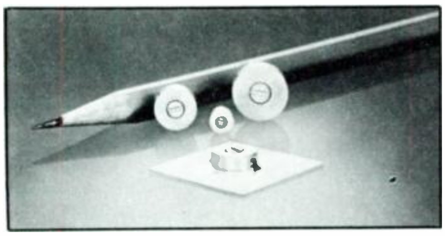
The FSC10FA/LF and FSC11FA/LF are low noise 0.5 μm gate length FET's designed for 4 GHz TVRO low noise amplifiers and general purpose gain stages to 6 GHz. The noise figure for FSC10FA/FSC10LF is NF = 0.7 dB typical, NF = 0.8 dB maximum and the FSC11FA/FSC11LF is NF = 1.2 dB maximum at 4 GHz with an associated gain, Ga = 13 dB. The linear gain is 18 dB and P1dB = 16 dBm for VDS = 3 V, IDS = 30 mA. These devices are available in two different packages: the LF is a low cost hermetic package and the FA is a .070 in square hermetic metal ceramic stripline package. **Fujitsu Microelectronics, Santa Clara, CA 95051 or INFO/CARD #173.**



INFO/CARD 77

New Capacitor Technology

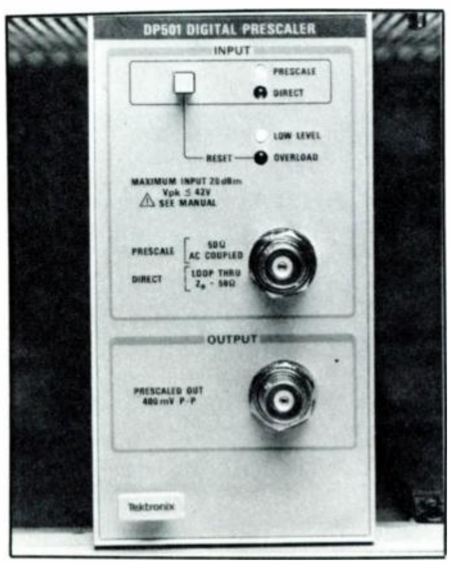
Two technological firsts in air variable capacitors have been announced by E.F. Johnson Company Components Division with the introduction of the Dyne Gap-SSM substrate mount capacitor. It is the first air variable capacitor which functions by varying the air gap between two metal plates, rather than by varying the inter-meshing of a series of metal plates. It also is the first capacitor to use the mounting surface as in integral part of the capacitor. This creates a capacitor of



extremely small size and low height (.122") making it ideal for hybrid and miniature circuits in the higher frequency range. The substrate mount Dyne Gap capacitor is designed for frequency trimming, filter tuning, impedance matching and remote control circuit tuning applications in hybrids and other miniature circuits. Designed for a wide range of frequency uses, its air dielectric provides high stability, and it features multi-turn precision tuning, silver plating on contacts, and extremely low inductance loses (.9 nano Henry). In the typical air dielectric capacitor, the capacitance is determined by varying the meshing of metal plates. The Dyne Gap, instead of varying the interface between multiple meshing plates, varies the air gap between only two plates. More important, the design permits the screening of the stator plate onto the substrate or PC board as an integral part of the circuitry. The Dyne Gap-SSM, with the rotor plate, is then solder-mounted over the screened stator to create a low profile, compact component. Since the design of the Dyne Gap capacitor calls for the stator to be fabricated by the user as part of the circuitry on the PC board or ceramic substrate, performance specifications are a function of the circuit board material and stator configuration used. However, performance specifications for two different circuit board materials are available to serve as a guideline. For a G10 PC board, typical performance specifications are 100 VDC working voltage, better than 10^5 Megohms insulation resistance and Q @ 1 MHz greater than 200. Temperature coefficient (measured with the capacitor set at 80% of its maximum capacity limit) is +400 ± 150 ppm/°C. Typical performance on alumina ceramic substrate is 100 VDC working voltage, greater than 10^5 Megohms insulation resistance and Q @ 1 MHz of greater than 2,000. Temperature coefficient in this example is −50 ± 150 ppm/°C. The Dyne Gap-SSM capacitor is available in three sizes: .500" diameter, .375" diameter and .219" diameter. All are just .122" high when mounted. Average selling price is $.50 per piece. **E.F. Johnson Company Components Division, Waseca, MN 56093 of INFO/CARD #176.**

Digital Prescaler

Tektronix has announced the introduction of the new DP 501 Digital Prescaler which adds 1.3-GHz frequency measurement capability to its TM 500 and TM 5000 digital counters, including most of those no longer manufactured. It can also be used, with some limitations, with non-



Tektronix counters, to extend their frequency range to as high as 1.3 GHz as well. When used with the Tektronix DC 5009 and DC 5010 GPIB Programmable Universal Counter/Timers, the DP 501 provides frequency measurement to 1.3 GHz under program control. When used with the Tektronix model 7L14 Spectrum Analyzer, TR 502 Tracking Generator, and a DC 509/5009 or DC 510/5010 counter, the DP 501 is able to provide the same "Bright Dot Marker" measurement of swept signals which has been available with the Tektronix model DC 508A Option 07 Digital Counter. The DP 501 is priced at $500, with four weeks delivery at the present time. **Tektronix, Inc. Beaverton, Oregon, 97075, or please circle INFO/CARD #175.**

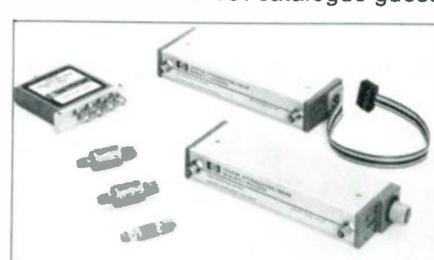
Microwave Components

Microwave system designers now have several new precision coaxial components available from Hewlett-Packard Company: a programmable 90-dB step attenuator; fixed attenuators, attenuators from 3 to 40 dB; and a novel coaxial-switch element for matrix switches. The HP 33323K is a coaxial step attenuator with 10 dB steps to 90 dB over the dc to the 26.5 GHz band. Key features include accuracy (+/−2.8 dB at 90 dB and 26.5 GHz) as well as low SWR (less than 1.8 at 26.5 GHz), combined with compact size and rugged design, qualify it for use in microwave-test systems. Three

models of fixed attenuators provide coverage from dc to 26.5 GHz. The HP 33340A operates from dc to 12 GHz; the HP 33340B from dc to 18 GHz, and the HP 33340C from dc to 2.5 GHz. In designing complex matrices for signal switching, the HP 33311B-CO4 coaxial switch provides a cross-bar switching action well-suited for microwave signals. It connects both crossing lines with low insertion loss or passes both signals through the intersection with at least 90 dB of isolation, dc to 18 GHz. By combining these switches with the common SPDT type, larger matrices may be configured. For example, a 3 × 4 matrix requires six HP 33311B-C04 switches and five of the SPDT type. The following are OEM prices for quantities 1-9: The HP 33323K is $1,785, with no charge for Options 008 and 016. The HP 33340A (12 GHz) is $85, the HP 33340B (18 GHz) is $110, the HP 33340C (26.5 GHz) is $238 and the HP 33311B-C04 is $735. **Hewlett-Packard Co. Palo Alto, CA 94303 or INFO/CARD #172.**

Two Filter Kits Aid Prototyping EMI Control

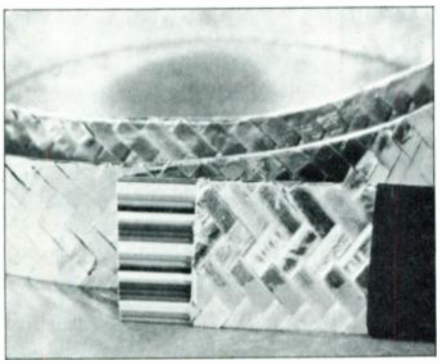
Two new filter kits allow electrical engineers to experiment with a variety of EMI filters without catalogue guess-



work. The contents of the kits can be installed in prototype devices to determine which are best suited for a specific application. One general purpose kit and one for digital systems are stocked with AMP Quiet Line filters and filtered assemblies. The filters, functioning between 2 MHz and 18 GHz, remove conducted electromagnetic interference above specified frequencies while producing no reflected noise. **AMP Incorporated, Harrisburg, PA 17105, or please circle INFO/CARD #124.**

Ribbon Cable Braiding

1983 F.C.C. RF shielding regulations can be achieved, utilizing an innovative ribbon cable shielding method perfected by Pacific Braid & Cable. This



new technique utilizes alluminized polyester, braided in 3/16" wide strips over flat ribbon cable. This production technique allows for long length interconnections to be RF shielded. Because a braid is utilized, full RF shielding is achieved without loss of cable flexibility! A PVC extruded jacket is used. **Pacific Braid & Cable, Huntington Beach, CA 92648, or please circle INFO/CARD #123.**

R.F. Millivoltmeter

A new r.f. millivoltmeter, the Model 9200A, is now available from Boonton Electronics. Basic accuracy is 1% of reading and calibrated measurements can be made from 10 kHz to 2.5 GHz. A high impedance probe and a full set of new accessories provide terminated thru-line measurements to 1.2 GHz in either 50-ohm or 75-ohm systems. An optional terminated voltage sensor extends 50-ohm measurements to 2.5 GHz. This sensor may also be used to display power levels to 2.5 GHz by selecting the dB display mode. A second input channel can be added to the 9200A. This option allows the simultaneous display of two voltage inputs, or their difference expressed in dB, and is ideal for measuring gain or loss directly. Display may be in mV, or dBm relative to any reference from 50 Ω to 600 Ω, dBr, dBmV, or

dBV, all with 0.01 dB resolution. Zeroing is automatic. A dc recorder output supplies a full-scale 10 volts that is linear with voltage over each decade voltage range, or linear in dB over the entire 80 dB range in any of the dB modes. A full function GPIB option, with selected termination characters and SRQ for efficient data transfer, adds complete talker/listener capability. Price: Model 9200A with high impedance Model 952001 Probe, $2595; optional 50-ohm terminated Model 952009 Sensor, $350; second input channel, $690; GPIB, $445. Availability: 6 weeks. **Boonton Electronics, Parsippany, NJ 07054, or INFO/CARD #122.**

Thin Film VCO's, 0.5 To 18 GHz

Technical specification sheets are available from Eaton Corporation's Communications Products Division describing a complete family of Addington VCO's covering the 0.5 GHz to 18 GHz bandwidths. These VCO's feature octave bandwidth fundamental oscillators up to 8 GHz and doubled frequency for RF outputs covering 18 GHz. **Eaton Corporation, Communications Products Division, Sunnyvale, CA or please circle INFO/CARD #121.**

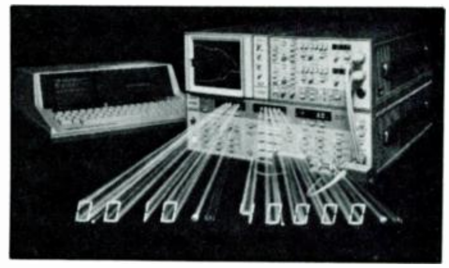
Signal Generator

The Texscan Psg 520 Synthesized Signal Generator is a compact lightweight unit covering the frequency range of 10-520 MHz. It is designed for laboratory or portable use and is capable of being powered from 110 VAC or 12 VDC. The low harmonic content and precise, 1 dB step, output control make it suitable for a wide

range of applications. Internal AM/FM modulation capability with digital readout of depth/deviation enables easy testing of both fixed and mobile receivers. A built in sinad measurement capability with digital readout further enhances the uses of the instrument. Reverse power protection of up to 50 W is a standard feature. Precise frequency setting in 100 Hz steps by means of thumb wheel switches means no resetting after power interruptions. RF leakage has been minimized to enable accurate measurements. External reference frequency and external modulation inputs add to the versatility of the PSG 520. **Texscan Corporation, Indianapolis, IN or please circle INFO/CARD #120.**

Coaxial Measurements From 10 MHz to 40 GHz

You can now make return loss, transmission loss or gain, and absolute power measurements from a single coaxial connector over the 10 MHz to 40 GHz range. Free from the constraints of cumbersome waveguide and multiple narrow-band test setups, you can build smaller, lighter devices with tighter specifications up to 40 GHz. Using the new WILTRON 2.92



mm "K" connector (SMA and 3.5 mm compatible), the system measures return loss with 30 dB directivity up to 40 GHz. Preprogrammed system software guides the user through step-by-step procedures with plain English menus and prompts. Residual system errors stored in memory are automatically subtracted from test data, which in turn are plotted as curves or printed in tabular form. All system elements cover the 10 MHz to 40 GHz range, making it unnecessary to interrupt tests to change plug-ins or to add fixtures. Option P2 locates faults up to 500 feet away along coax or waveguide transmission lines. Price: $57,135. Delivery: 12 weeks. **Wiltron Company, Mountain View, CA or please circle INFO/CARD #119.**

Spectrum Analyzer

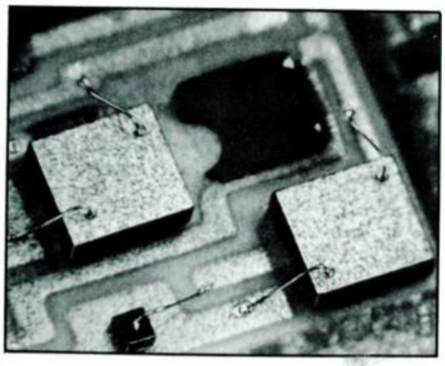
Investigation into audio and other vibration phenomena has taken a giant step forward with the introduction of the TECRON TEF[R] System 10, a microprocessor-based, two port test instrument. Don Eger, TECRON division manager stated that the TEF System 10 now offers great improvements in speed, convenience and portability for spectral analysis in the DC to 31 KHz range. The TEF System 10 is manufactured by the TECRON division of Crown International, Inc., Elkhart, IN, U.S.A., under license from Jet Propulsion Laboratories, California Institute of Technology, Pasadena, California. The TEF machine is an implementation of the research into swept-spectrum technology at JPL, and utilizes the principles of Time Delay Spectrometry (TDS) developed by Richard C. Heyser, a JPL staff member. The TEF System 10 comes complete in a 22" × 18" × 7.5" metal case, is sized to fit under airline seats

and weighs only forty pounds. It performs complete spectral analysis (DC to 31 KHz) of spaces, structures, components and materials. A complete analysis of a concert hall, for instance, can be accomplished with the TEF System 10 in much less than half a day. The major physical components of the TEF machine include a 92 character keyboard on a hinged cover, a 7" green-phosphor CRT, 5-1/4" minifloppy disk drive (dual disk optional), a rear panel with all input/output connections, three Z80 microprocessors and sealed lead/acid batteries which float on the DC bus, providing emergency power to complete tests in progress. The TEF System 10 is a 96K RAM computer, capable of performing all standard types of digital computation. It is a hybrid, utilizing both digital and analog signal processing. It is software based and menu driven, with a complete machine description and operating instructions built into software. Data-recording programs even include a "scratchpad" option to enter job descriptions. The TEF System 10 is offered standard equipment and software at $14,500.00. **TECRON division of Crown International, Inc., Elkhart, IN 46517, or INFO/CARD #118.**

Chip Resistors

Series CRC, a new line of commercial thick film chip resistors is now available from Dale Electronics, Inc. The new CRC chip resistors are designed for high density surface mounting in a wide range of commercial usage, including automotive, telecommunications and consumer applications. The new Dale chip resistors are being produced in the popular 3.2 mm × 1.6mm size, and are available with a choice of two terminations: Model CRC (terminations on the same side of the ceramic body as the resistor element); and Model CRCW (wraparound terminations). Both models have a nominal wattage rating of 1/8 watt at +70°C, and a continuous working voltage of 200 volts maximum.

They are available with a standard resistance range of 10 ohms to 2.2 megohms in ±2% and ±5% tolerances. Resistance temperature characteristic is ±200 ppm/°C over the operating temperature range of −55°C to +125°C. Packaging choices for the CRC series include tape and rell (EIA standard RS-481 with 4,000 pieces per reel) and bulk (5,000 pieces per plastic bag). Typical pricing for a CRCW chip resistor, wraparound termination, 5% tolerance, is 5 cents each in quantities of 5,000 pieces, bulk packaged. **Dale Electronics, Inc., Norfolk, NE 68701 or INFO/CARD #117.**



Single Layer Capacitor Chips

Johanson Dielectronics, Inc. announces a new series of parallel plate, single layer ceramic capacitor chips. With extremely high-Q characteristics and the low series inductance afforded by the single layer construction, these devices are ideally-suited for microwave applications ranging from D.C. blocking to coupling and bypass up to 50 GHz. Chip dimensions range from .020 squared to .200 squared with capacitance values from 0.1 pf to 16,800 pf. Voltage ratings of up to 15KV are available. Noble metal electrodes can be attached by either wire bondings, soldering or conductive epoxy techniques. **Johanson Dielectrics, Inc., Burbank, CA 91505 or INFO/CARD #116.**

High Speed Pulse Drivers

Colby Instruments, Inc., introduces four different models of very high-speed pulse drivers intended for high-speed digital electronics and com-

munication systems, and for general high-speed digital electronics and communication systems, and for general high-speed measurement applications. **Colby Instruments Inc., Los Angeles, CA 90073-0379 or please circle INFO/CARD #115.**

GaAs MES FET

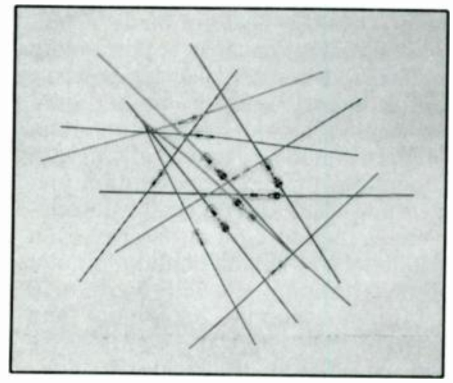
California Eastern Laboratories, Inc., introduces the NE41137, NEC Corporation's "new generation" GaAs MES FET. The NE41137 is a dual gate, GaAs N Channel MES FET designed for IF amplifier and mixer applications requiring Low Noise and high gain in the VHF and UHF bands. The NE41137 features a Noise Figure of 1.3 dB with an associated gain of 20 dB at 900 MHz with a low reverse capacitance of 0.03pf. The NE41137 is ideal for a wide variety of low noise applications through 2 GHz requiring reliability and low cost. Quality and reliability are assured by NEC's stringent quality control procedures. **California Eastern Laboratories, Inc., Santa Clara, CA, or INFO/CARD #114.**

RF Power Amplifiers

N.S.D. announces a complete line of RF Power Amplifiers for applications in Business, Broadcast & Military Communications. These amplifiers are available for use in laboratory instrumentation, mobile and base station installations and repeater relay systems. Frequency coverage is from 35 to 1000 MHz and output power to 500 Watts can be provided. Broadband techniques are utilized for adjustment free operation and models are available for both 12 and 24 VDC or 110/220 VAC operation. **N.S.D., Los Alamitos, CA or INFO/CARD #113.**

Current-Controlled Attenuator Diodes

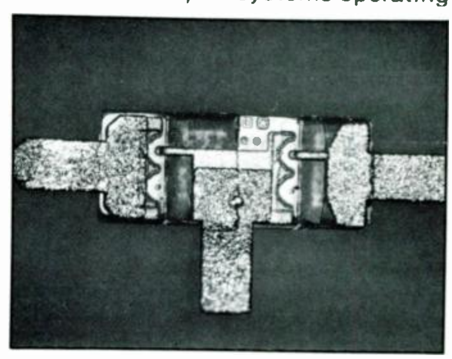
Alpha Industries has introduced a new series of Current Controlled Attenuator Diodes (DSB 6419 Series), specifically designed for variable resistance over a wide range of current. Variation ranges from 10,000 ohms (or more) at the low current end . . . up to between 2 and 4 ohms, depending on the device type. Manufactured by Alpha's Semiconductor Division, these diodes are designed



for use in current-controlled attenuator and modulator applications. They are also useful in applications involving electronically tuned filters, phase shifters, limiters, and switches. These devices have a long life time and a thick intrinsic region, allowing them to be used at frequencies as low as 1 MHz with low distortion. Several levels of resistance are available in the series. **Alpha Industries, Inc., Woburn, MA 01801, or INFO/CARD #112.**

Schottky Beam-Lead Pairs

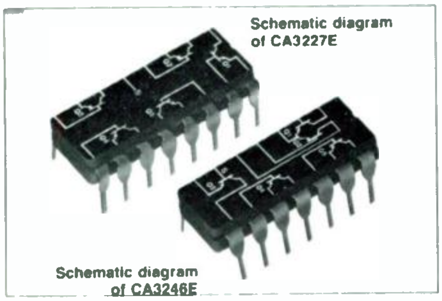
Microwave Schottky beam leads in pair configuration with a guaranteed maximum capacitance of 0.10 pf are now offered by Hewlett-Packard Company. HP officials believe Hewlett-Packard is the first manufacturer to offer silicon microwave Schottky diodes with these specifications. The low capacitance enables the designer to achieve a low-noise figure (7.0 dB max. at 16 GHz) for systems operating



at high frequencies. **Hewlett-Packard, Palo Alto, CA 94304, or INFO/CARD #111.**

NPN Transistor Arrays

Two high frequency NPN transistor-array integrated circuits from the RCA Solid State Division are designed for applications at frequencies up to 1.5 GHz. Each array consists of five general purpose silicon NPN transistors on a common monolithic substrate. The CA3246E array consists of three independent transistors plus two differentially connected devices housed in a 14-lead dual-in-line plastic package. The CA3227E consists of five independent transistors in a 16-lead plastic dual-in-line package. The CA 3246E and the CA3227E are high frequency versions of the popular CA3046 and CA 3127E arrays, respectively.


Schematic diagram of CA3227E

Schematic diagram of CA3246E

Both IC arrays operate over the full military temperature range of −55 to +125°C. Monolithic construction provides close thermal and electrical matching of the five transistors. In quantities of 1000 pieces, the CA 3246E is priced at $1.93 and the CA 3227E at $2.18 (U.S. only). **RCA Solid State Division, Somerville, NJ 08876, or INFO/CARD #110.**

ECL Crystal Oscillator

Model E14R1 drives 100k and 10k ECL at any fixed frequency from 25MHz to 200 MHz with total frequency tolerance of ±.01% over 0°C to +70°C, symmetry 50 ± 10%, risetime and falltime ≤1.5ns, supply voltage −4.5Vdc

to −5.2Vdc ± 5%, supply current <50 ma; all metal 4 pin package measures 0.8" (20.3mm) L × 0.5" (12.7mm) W × 0.2" (6.4mm) H has complimentary output. **The Connor-Winfield Corporation, West Chicago, IL 60185 or INFO/CARD #109.**

New Literature

Microwave Semiconductor Catalog

Now available from California Eastern Laboratories, Inc., the new 48-page catalog covering NEC's complete line of microwave semiconductors. Included are low noise/small signal and power GaAs FETs, small signal and power bipolar transistors, diodes, and components, featuring silicon monolithic and hybrid amplifiers. Performance curves, specifications and detailed package outline drawings are provided. The catalog is conveniently divided into small signal, medium power and power categories. **California Eastern Laboratories, Inc., Santa Clara, CA 95050 or INFO/CARD #127.**

Reconditioned Test Equipment and Components

A new 238 page catalog published by Tucker Electronics Company lists approximately 4100 different pieces of reconditioned electronic test equipment and microwave components. Instrument categories include: amplifiers, analyzers, bridges, frequency measuring equipment, signal generators, lab standards, meters, scopes, power supplies, recorders, RFI/EMI equipment and more. Each unit is described and priced. All instruments available for either *sale* or short-term *rental*. All units are reconditioned and calibrated to manufacturer's specifications. **Tucker Electronics Company, Garland, Texas 75042, or INFO/CARD #126.**

Video and T.V. Components Catalog

Marshall Electronics Inc. of Culver City, CA is now releasing their 1983 catalog for the Video, Audio and Satellite industries. The catalog shows a wide range of high quality Video, TV and Audio components for professional, industrial usage. The items are used in virtually every major area of Audio/Video communications including: Broadcast, CCTV, CATV, MATV, Satellite and Sound Recording. New items high-lighted in the catalog are: High frequency electronic switching devices for 54 MHz-5 GHz operation. Satellite 4 and 8 way splitters. 'N', connectors for Satellite use and ultra miniature resonator and crystal controlled RF modulators. A complete range of connectors included in the catalog are BNC, 'N', 'F', RCA, and multi-plan push lock. **Marshall Electronics Inc., Culver City, CA 90230 or INFO/CARD #125.**

RF Amplifier Catalogue

The new short form catalogue from Q-Bit Corporation describes a range of standard rf amplifiers that operate within the overall frequency range 0.5 to 1200 MHz. The impedance match and inherently high reverse isolation of the amplifiers facilitates their operation in cascade without degrading the performance of other stages and the catalogue contains details of packages that are configured with up to 5 units. The range consists of hybrid (designated QBH) and modular (QB) series, thus the catalogue tabulates the electrical characteristics of approximately 50 units, offering a wide choice of noise figure, gain and power capability throughout the frequency range. Single hybrid units are supplied in TO-8 housing or pin-connected modular packages for multi-stage operation. Dimensioned outline drawings are included for all units and a choice of connectors is offered for most connectorized modules. All devices are quality screened during manufacture and details of screening levels (up to MIL standard specification) are quoted in the catalogue. **March Microwave Limited, Braintree, Essex or INFO/CARD #137.**

Planar Tubes Brochure

A new 8-page, color brochure from the Special Products Division of siemans Components, Inc., outlines the company's extensive planar tube product line. Siemens provides a wide range of planar tubes for TV applications, pulse operation, RF amplifiers and oscillators. The brochure catalogs the various types according to application, listing type number and technical specifications as well as ordering information. **Siemens Components, Inc., Iselin, NJ 08830, or INFO/CARD #139.**

Capacitor Catalog

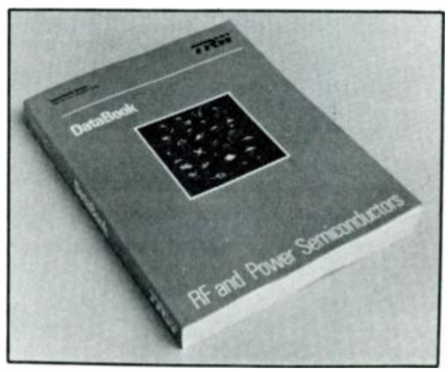
A new 40-page Catalog/Technical Manual providing detailed information on the application and performance of fixed ceramic, variable ceramic/air capacitors through Microwave frequencies is now available from Murata Erie. This new manual includes detailed performance curves for Q, self resonant frequency, drift, temperature and more for both fixed and variable units. Also included are detailed test and performance instrumentation set-up parameters, a glossary of tems and catalog information on the complete Murata Erie line of high frequency, high Q devices. **Murata Erie North America, Inc., Marietta, GA or INFO/CARD #138.**

Glass Sealed Electronic Connectors

Four page short form catalog and cross reference on Glass Sealed Hermetic Electronic Connectors. Provides ILLUSTRATIONS OF VARIOUS MIL-SPEC Style Connectors, Materials and Finishes. **Detoronics Corporation, So. El Monte, CA 91773 or INFO/CARD #136.**

Semiconductor Data Book

The Semiconductor Division of TRW Electronics Components Group today announced the immediate availability of its newest and most comprehensive data book ever on *RF and Power Semiconductors.* The data book contains over 600 pages of product photos, complete electrical specifications, and



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parametric performance characteristics on 360 RF and power devices. Products listed range from LVA Zener and Shottky diodes to broadband microwave transistor to RF linear hybrid amplifiers. A leader in semiconductor technology for many years, the Semiconductor Division manufactures devices in 18 different product categories. Some of the applications for TRW Semiconductor devices are CATV, mobile and aircraft radio, missile telemetry, pulse radar systems, switching power supplies, and microwave transmitters and receivers. **Semiconductor Division, TRW Electronic Components Group, Lawndale, CA 90260, or INFO/CARD #135.**

MLC Chip Capacitors, Inductors and Transformers Brochure

An expanded line of leadless ceramic chip capacitors and chip inductors/transformers is described in a 28-page catalog from SFE Technologies, San Fernando, California. The illustrated publication devotes several pages to a description of the technical characteristics of its products, including voltage-temperature parameters, aging, voltage effects, terminations, component attachment, performance/environmental testing, packaging and marking. **SFE Technologies, San Fernando, CA 91341 or INFO/CARD #134.**

Transmitting and Generator Tubes Brochure

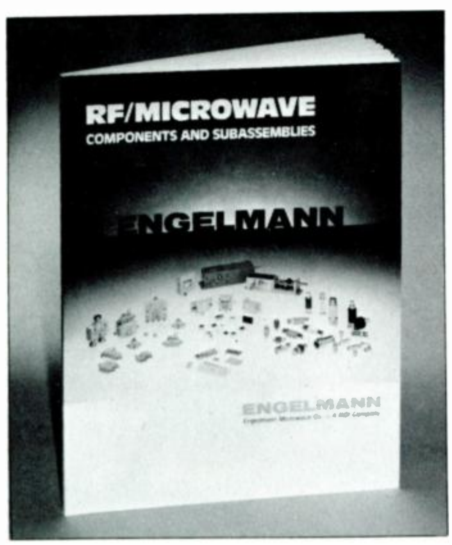
The Special Products division of Siemens Components, Inc. has just issued a 16-page brochure covering the company's line of transmitting and generator tubes. The brochure provides a clear survey of the current product line listing the tubes by main application. Within the individual application fields, the tubes are divided according to power classes based on the most common operating characteristics. Also contained is a summary of replacement tubes and a list of tubes in alpha-numerical order. **Siemens Components, Inc., Iselin, NJ 08830, or INFO/CARD #133.**

RF Test Equipment Catalog

A new 22-page High Power RF Instruments catalog of coaxial Load Resistors, rigid-line directional THRU-LINE© Wattmeters, precision Calorimeters, WATTCHER© Power Monitor/Alarms and RF Filters is available from Bird Electronic Corporation. This comprehensive reference of High-Power and Broadcast RF measurement instrumentation from 250 watts to 250 kilowatts and from 2 to 1000MHz also includes self-cooled, water-cooled and air-cooled loads (to 2GHz) and a Tech Data section with rigid transmission line data. **Bird Electronic Corporation, Cleveland, Solon, OH 44139, or INFO/CARD #132.**

Master Catalog

Engelmann Microwave, a KDI Company has just released a new 80 page master catalog on its full line of RF/MICROWAVE COMPONENTS AND



SUBASSEMBLIES. Included in this master catalog are tutorial sections on attenuators describing how they are used, important specifications, accuracy and frequency response, power and temperature sensitivity, schematics and pull-out drawings of each basic type. The tutorial introduction to couplers include descriptions of important specifications the designer should know like transmission loss, insertion loss, directivity, frequency sensitivity. This section also includes schematics illustrating every type of coupler available. The Englemann Catalog gives full specifications, schematics, photographs and outline drawings on its complete line of RF/Microwave Components and Subassemblies including: attenuators, hybrid couplers, directional couplers, power dividers/combiners, power monitors, blocks & bias tees, detectors, filters, toroidal mixers, oscillators, phase shifters, diode switches, terminations, and tuners. **Englemann Microwave Co., Whippany, NJ 07981, or INFO/CARD #131.**

Product Catalog

The Catalog is a 56 page Catalog with a complete coax cable section. Also, included is a competitive cross

reference index, interface drawings per MIL-C-39012 & information on SMA, TNC, Precision N & Hi-Voltage connectors & between series adapters. **United Microwave Products, Inc., Torrance, CA 90503, or INFO/CARD #130.**

904/907A Product Brochure

Two Microwave Signal Generators covering 3.7 to 12.4 GHz are described

in a new 6-page color brochure from Wavetek. This brochure identifies their unique design characteristics and operator benefits, as well as their suitability for radar, communications, component test and education applications. The all-solid-state Models 904 and 907A cover 3.7 to 7.6 GHz and 7.0 to 12.4 GHz, respectively, with calibrated power from 0 to −127 dBm. FM, AM pulse modulation, full band sweep and external voltage control are standard. GPIB programming is optional. **Wavetek San Diego, Inc., San Diego, CA 92138, or INFO/CARD #129.**

Panel Meter and Test Instrument Catalog

Simpson Electric Company is offering their new free 68-page, four-color Catalog 5300, listing their complete line of stock analog and digital panel meters, meter relays, controllers, sound level meters and test instruments. **Simpson Electric Co., Elgin, IL 60120, or INFO/CARD #128.**

New Books

Digital PLL Frequency Synthesizers Theory and Design

By Dr. Ulrich L. Rohde
Prentice-Hall, NJ, 1983

Dr. Rohde's book is a welcome addition to the formal PLL literature. Many of the present books are too much influenced by the old servo system designs. Dr. Rohde takes a more modern approach, both in theory and applications. The book contains the description of the more realistic and useful third order loop as well as other less known configurations. Dr. Rohde had the advantage of having access to the Rohde & Schwartz PLL designs. Thus, his examples are actual circuits proven in successful commercial designs. The circuits are fairly recent, as books go, since some of the designs are less than two years old. By the same token, the components used to illustrate the circuitry are still commercially available and include some of the latest integrated

HP 8684B 5.4-12.5 GHz; AM, FM, Pulse; +10 to −130 dBm.

HP 8683A 2.3-6.5 GHz; AM, FM;
0 to −130 dBm.

Diagnostics can be displayed on
a remote CRT terminal.

ENVIRONMENTALLY TYPE-TESTED

Temperature: Operating: 0 to +55°C
Non-operating: −40 to +65°C
Humidity: 95% RH at 55°C
Shock: 30 g at 11 ms; 18 shocks
Vibration: 5 to 25 Hz at 0.025" p-p
25 to 55 Hz at 0.020" p-p
Altitude: Operating: 15,000 ft.
EMI: MIL-STD-461A

High reliability is no accident in HP's 8683/84 cavity-tuned signal generators.

Designed for MIL-T-28800B, Class IV environments.

Every component in the 8683/84 is stress analyzed under HP's thorough engineering reliability program. For instance, even at 50° C ambient, junction temperatures of semiconductors remain below 50% of manufacturer ratings. Internal thermal design included infrared scanning to pinpoint hot spots for correction.

Sub-modules as well as complete generators were type-tested to the rigorous MIL-T-28800B Class IV environmental standard. The result: rugged, portable signal generators that ensure high performance signal simulation and spectral purity.

6000 hours of demonstrated MTBF

In addition to the above military standard tests, HP subjected 12 production units to accelerated high-temperature life tests, demonstrating a mean-time-before-failure better than 6000 hours (with a 90% confidence factor). On average, that's three years of trouble-free operation! However, based on HP's first-year warranty reports, actual reported MTBF exceeds 10,000 hours. That gives you real confidence in the field or in the lab.

Built-in-test serviceability ensures low ownership cost.

Each 8683/84 features built-in-test capability, controlled by the internal microprocessor. You get front panel diagnostics to display more than 38 internal tests. Plus, you can connect a terminal to a built-in RS-232 port or have a telephone/modem link to a central location for evaluation by an experienced service engineer.

Communications Applications	Radar/EW Applications
8683A .... 2.3-6.5 GHz	8683B .... 2.3-6.5 GHz
8684A .... 5.4-12.5 GHz	8684B .... 5.4-12.5 GHz
Select the 8683A and 8684A models for communications field applications: wide band FM to 10 MHz, metered AM, non-harmonic spurious <−85 dBc. $12,000*	Use the B-models when you need AM, FM, plus internal pulse generator and modulator with >80 dB on/off ratio and <10 ns rise/fall times. $15,000*

Find out more about HP's rugged, 38 lb., 8683/84 Signal Generators by sending for a free data sheet and product note. Just call your nearby HP Sales Office, or write: Hewlett-Packard, 1820 Embarcadero Rd., Palo Alto, CA 94303.

0401302A

*U.S. domestic prices only.

HEWLETT PACKARD

INFO/CARD 89

circuits. The author provides many practical design hints and actual performance data. Some of the phase noise information, for instance, is not readily available in books and is difficult to find in periodicals. In general, the book is a good practical guide to modern PLL design. While providing the necessary theoretical background, some of which could be found in other books, it provides the designer with a lot of practical information to design successful PLL circuits. And, it provides all this information in one reference. Speaking of references, many diverse references are cited. Not only the standard classical texts, but also the usually ignored amateur radio applications and articles in popular periodicals. Another welcome addition, for some designers, is the inclusion of computer programs written in BASIC. Nothing is absolutely perfect, and this book has some flaws. For instance, a common typographical error is the use of "+" for ":" in some of the programs and figures. These can be usually easily spotted, however. It is hoped that future editions will have these errors corrected.

All symbols are thoroughly defined at the beginning of the book. This is very important, since different authors use different notations.

The book is divided into 6 chapters, plus an appendix. The first chapter covers general theory of first, second and third order loops of both type 1 and 2. The second chapter contains phase noise information not found in other reference books. Additional phase noise information can also be found in Chapter 5, which describes different PLL systems. Chapter 3 describes several possible PLL configurations. These include multi-loop and fractional-N methods. Chapter 4 is also quite useful, since it describes, in considerable detail, modern circuits suitable for the individual PLL components. These include not only the theoretical design information, but also actual examples and their performance. Chapter 6 gives schematics of actual synthesizer designs. These are in addition to the applicable examples given throughout the book.

The Appendix consists of mathematical review and calculator/computer programs. The math includes such subjects as the Laplace transforms and manipulations of complex variables, but again stresses practical design by giving actual circuit examples. The calculator/computer programs provide the means for a complete analysis of the more complicated loops.

In general, this book, more than its predecessors, will be useful to the designer, mainly because of the modern circuit examples and actual circuit performance data.

– Andrzej B. Przedpelski
A.R.F. Products, Inc.
R&D Laboratory
Boulder, Colorado

Available from Prentice-Hall Inc., Englewood Cliffs, NJ 07632, 608 pages, $49.95, Hard cover, first edition 1983.

70 Years of Radio Tubes and Valves

By John W. Stokes

The entire electronics industry was founded on radio's "Magic Lamp" — the radio 'tube,' or 'valve,' in British parlance. From its invention in 1904 to its gradual eclipse beginning in the 1960's as it was replaced by solid-state devices, the electron tube reigned supreme as the cornerstone of the electronics industry.

In this new book the history of the radio tube is covered in its entirety, with emphasis on developments occurring between 1927 and 1937 — the period when the 'all electric' receiver evolved to become a familiar part of our daily lives.

Mr. Stokes, a New Zealand native, has been a student of tubes and valves for many years, and he gives equal treatment to developments throughout the Western World. With over 430 pictures, drawings, and early advertisements, the book is a treat for anyone involved in electronics to see and read how this nōw-giant industry which permeates every aspect of today's existence grew together with inventions in the field of tubes.

All the giants of the industry — Westinghouse, General Electric, Sylvania, RCA, Raytheon and others their size were involved, as were many smaller firms — a lot of them forgotten today, as if awaiting a book like this to revive their memories for anyone who enjoys studying how our modern technical civilization has progressed. The story is thorough, from Edison's discovery that electrons would flow in a vacuum (the famed 'Edison effect') to Lee DeForest's invention of the 'grid,' to RCA's 'Nuvistor' that closed the era.

With an estimated 8000 antique radio buffs in the USA alone, together with some 200,000 'Ham' radio operators, there's an enormous interest in old radio today. Mr. Stokes' book is intended to serve this market, as well as today's electronic engineers who wish to learn more about the rich heritage of their industry.

Available from The Vestal Press Ltd., Vestal, N.Y. 13850, 256 pages, $12.95, Hardcover, 1982.

Crystal Oscillator Circuits

By Robert J. Matthys

The performance of a crystal oscillator is highly dependent on its circuit design, and many different types of circuits are in use — each with its good and bad points. Some work better than others; some are designed for low cost and some for maximum stability; others to interface with digital logic circuitry.

This book responds to the need to analyze all types of solid-state crystal oscillators according to how they work, how to design them, and which ones work best.

Both fundamental oscillators from 1 KHz to 20 MHz and VHF harmonic oscillators from 20 to 100 MHz are covered. Many new circuits that have not appeared in publications before are included.

Contents: background, quartz crystals, fundamentals of crystal oscillation, circuit design characteristics, basic oscillator circuits, short-term frequency stability, analysis of phase versus frequency, trimming the crystal frequency, the start-up problem, discrete transistor oscillators, integrated circuit oscillators, comparing the circuits, optimizing the oscillator circuit.

Available from John Wiley & Sons, Somerset, NJ 08873, 233 pages, Hardcover, 1983.

RF Circuit Design

By Chris Bowick

"RF Circuit Design" is written for those who desire a practical approach to the design of rf amplifiers, impedance matching networks, and filters. If you are an individual who has little rf circuit design experience, you can use this book as a catalog of circuits, using component values designed for your application. On the other hand, if you are interested in the theory behind the rf circuitry being designed, you can use the more detailed information that is provided for in-depth study. **Contents:** Components, resonant circuits, filter design, impedance matching, the transistor at radio frequencies, small-signal rf amplifier design, rf power amplifiers, vector algebra, noise calculations.

Available from Howard W. Sams & Co. Inc., Indianapolis, IN 46268, 176 pages, $22.95, Softcover, 1982. ☐

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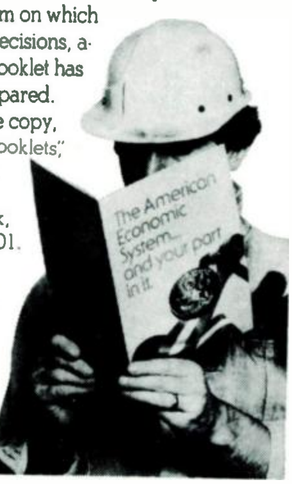

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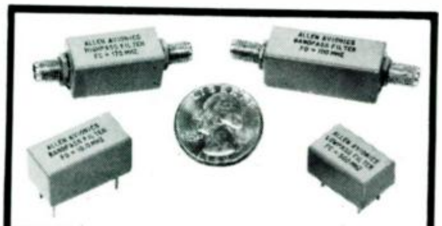

Sub-Miniature L-C FILTERS

ALLEN AVIONICS Sub-Miniature Passive L-C Lowpass, Highpass and Bandpass Filters range from 5 MHz to 500 MHz. The filters are offered in a 50 ohm impedance with SMA connectors or terminals for PC mounting. Several different shape factors are available. Units range in size from .8 x .5 x .5 to 1.25 x .5 x .5 inches.

PHONE/WRITE FOR FILTERS AND DELAY LINES CATALOGS

ALLEN AVIONICS, INC.
224 East 2nd St., Mineola, NY 11501
Phone: 516-248-8080

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ADVERTISER INDEX

Now is the time to rethink your design decisions—if you require up to 1 watt output for low-distortion intermodulation testing . . . broadband isolation . . . flat gain over wide bandwidth—or if you need much higher output level from your signal/sweep generator or frequency synthesizer — you can now specify Mini-Circuits' new ZHL-42 power amplifier . . . for only $895.

Using ultra-linear Class A design, this state-of-the-art four-stage amplifier provides 30dB gain, flat (±1.0dB) over the 700 to 4200 MHz range, is unconditionally stable, includes overvoltage protection, and can be connected to any load impedance without amplifier damage or oscillation.

One week delivery . . . and, of course, one-year guarantee.

ZHL-42 SPECIFICATIONS

Frequency	.7 to 4.2 GHz
Gain	30dB Min.
Gain Flatness	±1.0dB
Power Out @ 1dB CP	+29dBm Min.
VSWR In/Out	2.0:1 Max.
Noise Figure	7.5dB
Supply	+15V @ 690mA
Third Order Intercept	38 dBm Min.
Second Order Intercept	48 dBm Min.
Size	7" × 3¼" × 2⅛"h

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setting higher standards

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