March/April 1984

A Cardiff Publication

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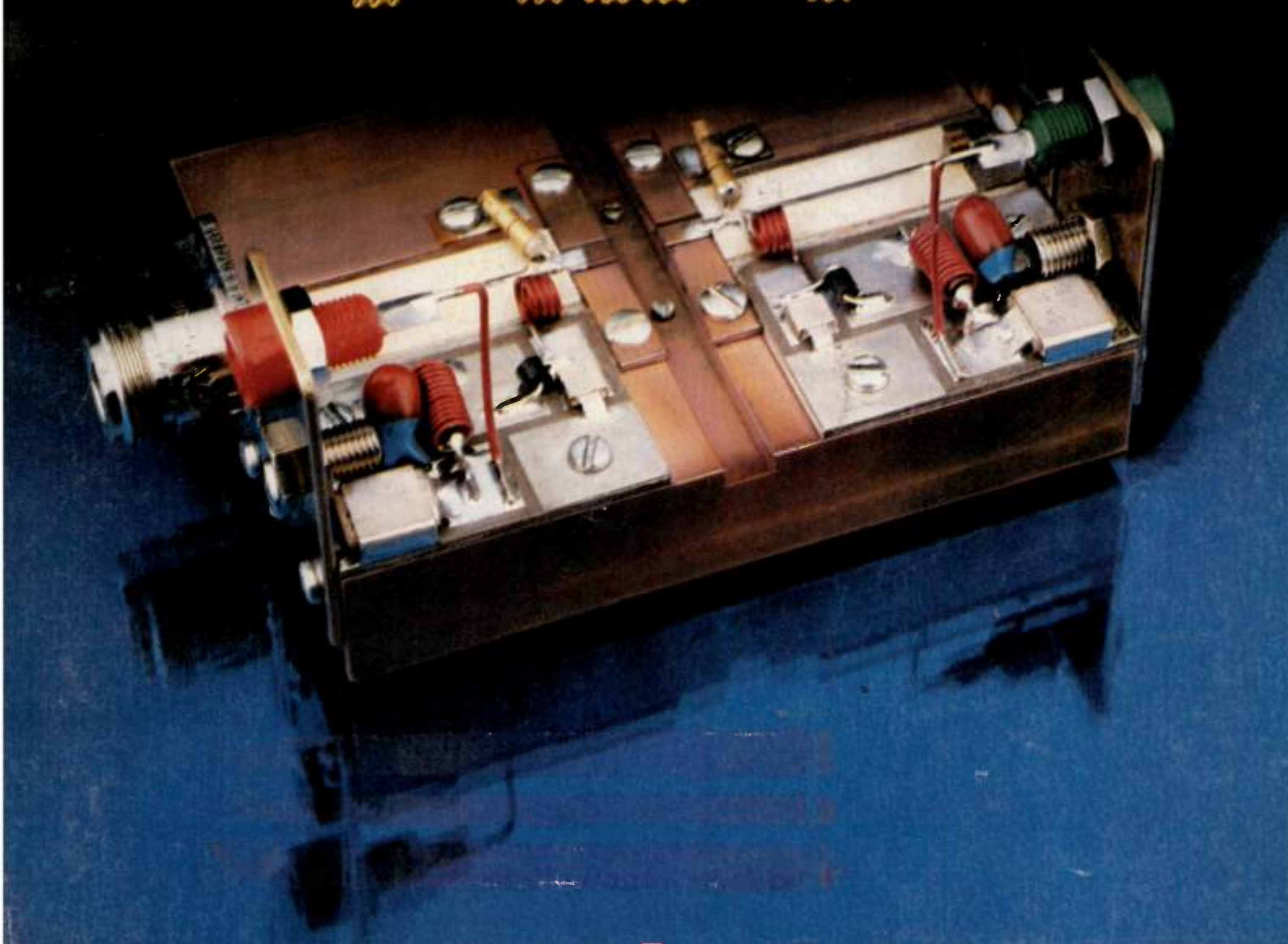

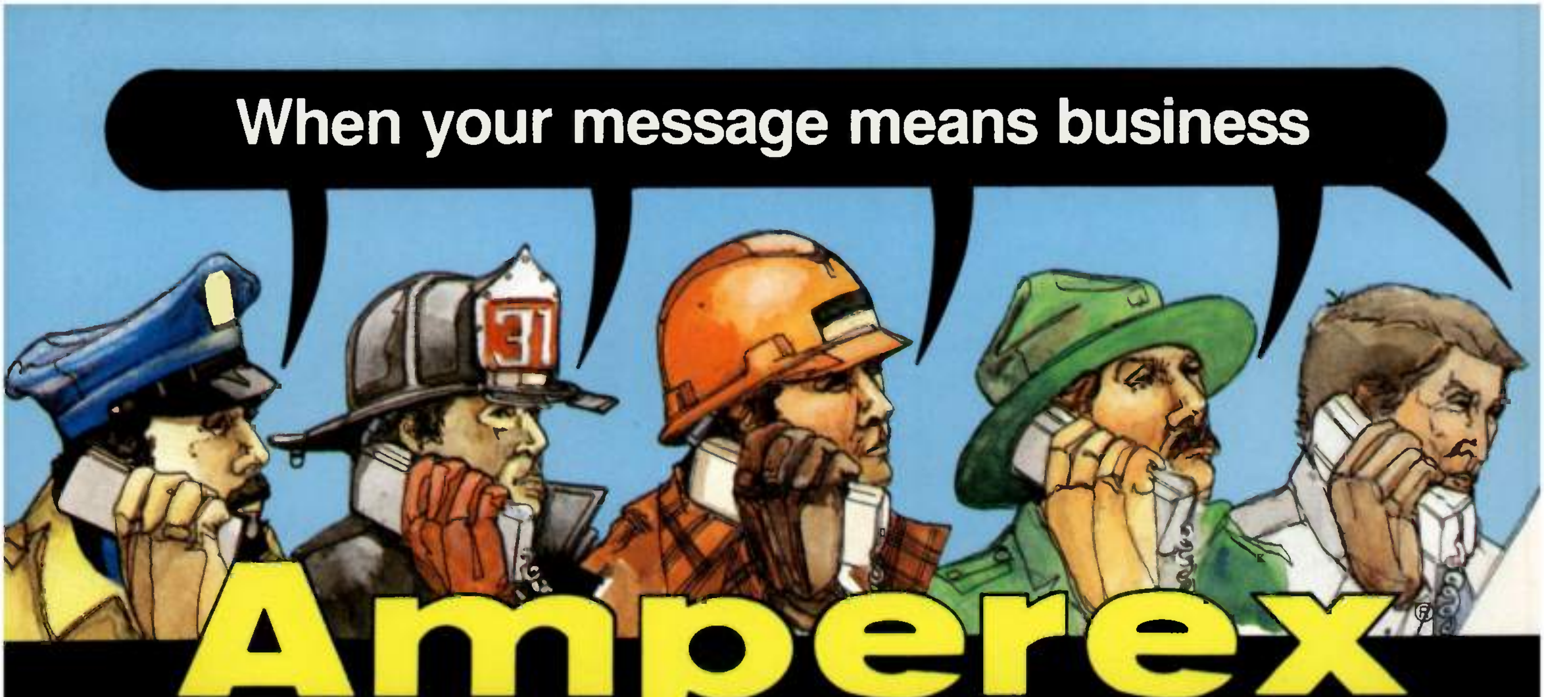

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BLV90	1	900	7.5	12.5	SOT-172
BLV91	2	900	6.5	12.5	SOT-172
BLU99	4	900	7.3	12.5	SOT-122
BLV92	4	900	8.0	12.5	SOT-171
BLV93	8	900	6.0	12.5	SOT-171
BLV94	15	900	6.0	12.5	SOT-171

400 to 512 MHz	MOBILE APPLICATIONS				
BLU60/12	60	470	4.8	12.5	SOT-119
BLU45/12	45	470	5.1	12.5	SOT-119
BLU30/12	30	470	6.0	12.5	SOT-119
BLU20/12	20	470	6.5	12.5	SOT-119
BLW82	30	470	5.0	12.5	SOT-119
BLW81	10	470	6.0	12.5	SOT-122
BLU99	5	470	10.5	12.5	SOT-122
BLW80	4	470	8.0	12.5	SOT-122
BLW79	2	470	9.0	12.5	SOT-122
BLX65	2	470	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
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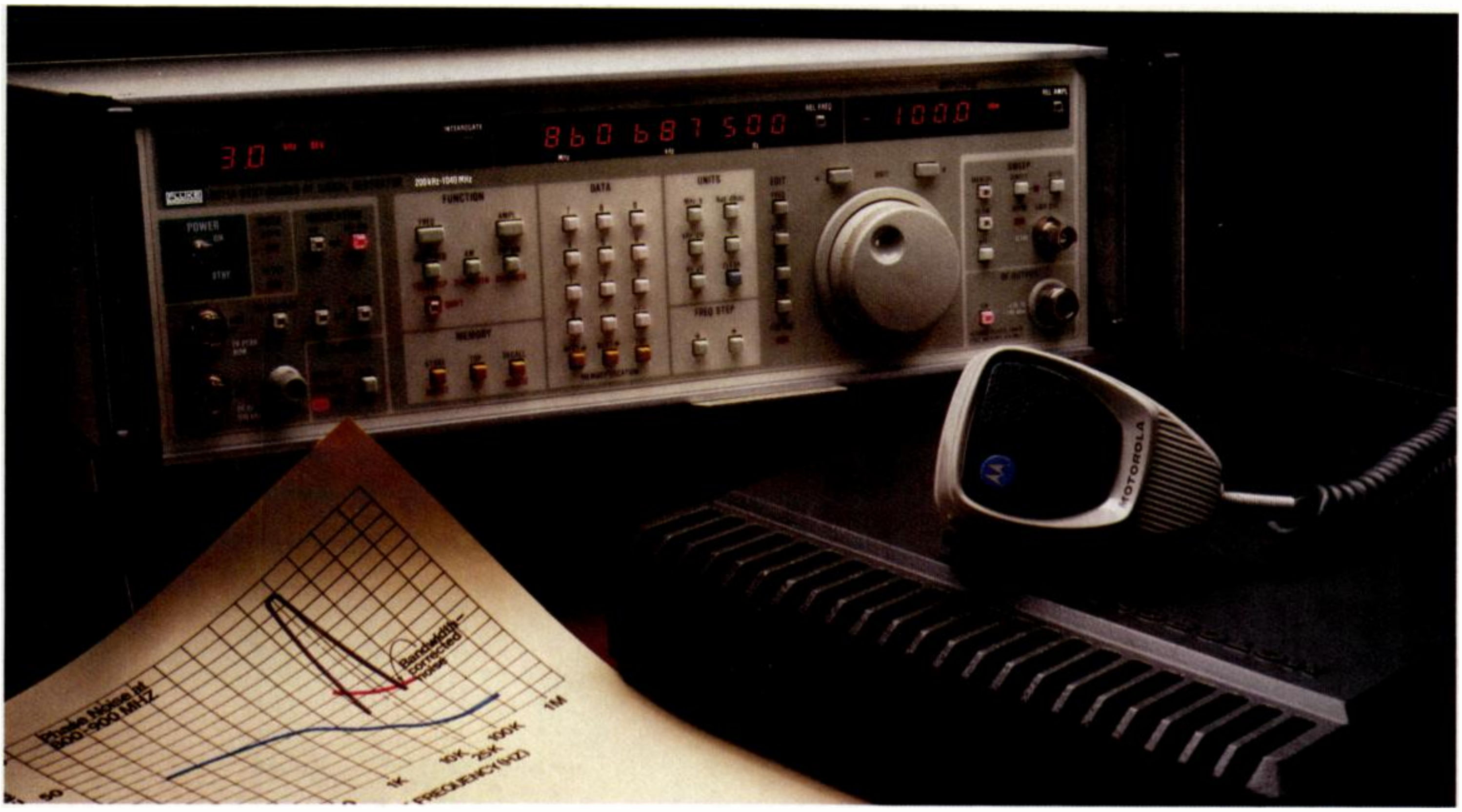

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INFO/CARD 2

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March/April 1984

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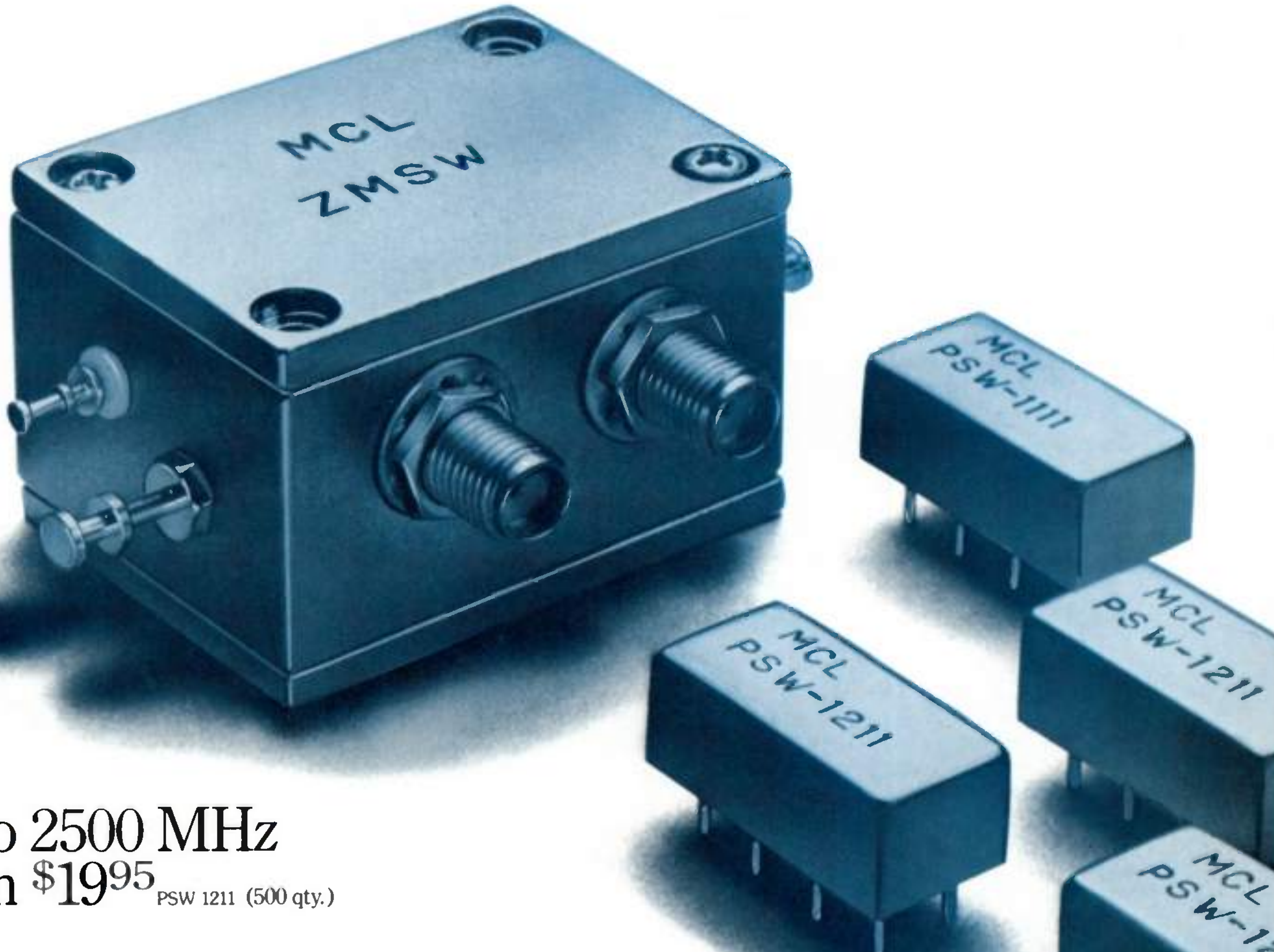

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INFO/CARD 3

"Basic, The Universal Language?" Part III

Dear Editor:

Mr. O'Neil (Sept./Oct. 1983) has indeed sparked a controversy! Sides are forming as loyal defenders of the faith are rushing to the battlements to defend the Holy Grail; in this case either the programmable calculator or the personal computer. As one who has walked down both paths the last 16 years, I would like to share a few comparative thoughts.

The programmable calculator and the personal computer are as alike, and as different, as the tack hammer and the sledge; each is a tool and each has its uses.

I have some 44 programs that I regularly use on my HP-41CV programmable system. In terms of size, the average program takes up 57 storage registers. One is only 10 registers long while another is 296; the short one returns physical line length as a function of frequency and dielectric constant, while the long one will tell you almost everything you could want to know about all types of microwave bandpass and bandstop filters and exactly how to fabricate each one. In between, almost every conceivable eventuality is covered: one determines the Fano bandwidth limits, another fully implements the Smith Chart in both lumped and distributed form, still others cover all forms of single and coupled line calculations.

I can use my programmable at my desk, on the bench, in a car, or on business trips all thanks to its portability and battery power source. And, because of the extensive built-in math functions, I can quickly program a solution to most any design problem. I do not have to write a subroutine to make a simple polar to rectangular conversion! I can even use it at the supermarket to add up the groceries.

On the other hand, my personal computer is faster and has a good deal more memory. However, these are relative terms in actuality since my average program is so short. Who cares if I can shave off a few tenths of a second in calculating a line length? And who really runs programs requiring 5 to 64K of memory everyday? I use my PC for what it is best at, word processing, spread sheet, and generating reams of data on microstrip, stripline, and coupler characteristics; I have 5 programs that do all of that. I also run one program employing a gradient search adaptive algorithm for wide band amplifier design. It needs the speed of my PC and runs for hours; it is so big it takes up 40 feet of printer paper!

I only use my personal computer in my office. I don't like to carry it around. I use it for big things. On occasion I even take command of the Enterprise and go out to fight off the Klingons!

I think the point has been made? The majority of the calculations made by the average RF/Microwave Engineer are relatively short and can be handled quite well on a little handheld that you can slip into your shirt pocket. We mostly deal with "tack hammer" type tasks. On the other hand, the generation of tabular data, the optimization of large networks, or the creation of ones own in-house version of COMPACT are all "sledge hammer" tasks and make good use of the speed and memory of a personal computer.

Perhaps something that would make everyone happy would be for a young and knowledgeable engineer with time on his/her hands to just sit down and write an article detailing a translator program? Translator routines exist in the larger PC world, why not a routine that could simply and quickly turn an HP or TI program into Apple, Commodore, IBM, or Radio Shack Basic? Is anyone out there listening? If you have been waiting for an idea, something that many would appreciate and use, something that r.f. design would snap up and publish in an instant, get off the dime and GO FOR IT!

Sincerely,

James J. Lev
Mgr. RF/Microwave Design
Cartwright Engineering Inc.

Dear Editor,

Should "Basic" be the universal language. I do not think it should. I am presently converting a Fortran 4 program published in the April 1980 r.f. design, page 28, Design of a Miniature Quadrature coupler to use on my TI59. I feel you have published interesting solutions in the past and programs that I can use on calculators I can carry in and out of the plant. We do have the large circuit design systems here at Hughes but they are not always available to the working circuit designer, when he needs them. Most circuit men I know like their handheld programmable calculators, and I have yet to see the personal computer run a Fortran 4 program with complex calculations.

Sincerely,

Robert Gunderson
Hughes Aircraft Co.

Dear Editor:

I have been reading with some amusement the BASIC versus Calculator "controversy." It seems to me that any well written article would not have this problem because it would include the kind of complete documentation one would expect of an engineer, namely: all formulas used, definition of all variables, and an example or two for reference.

Perhaps proponents of each tool (for computer and calculators are merely tools for an RF Engineer) could provide an article on "Translating BASIC to Calculators" and "BASIC Translations of Calculator Subroutines and RPN," especially in the Rectangular to Polar, Radians to Degrees, and Stack usage areas. There are differences in Labeling and use of subroutine calling, etc. as well.

I use both an HP-25 and Timex 1000 in my work . . . the Timex for formatted calculations with hard copy, the HP calculator for general simple calculalations. Each has its strengths and its weaknesses. Both are equally useful depending on application.

Sincerely,

James Eagleson
R.F. Engineer
Identronix, Inc.

Dear Editor,

I would like to congratulate you on the improvement in the quality of the articles in RF Design in the last year. Many of them have been very useful for some of my recent projects.

The recent letters on the usefulness of calculator programs versus Basic programs for home computers were interesting. As an unbiased observer, I find it hard to beat the usefulness of a programmable calculator for engineering applications. Besides that, most of the programs for the calculator are free while the computer programs are over priced in many instances. Also, most computer programs are not well documented be-

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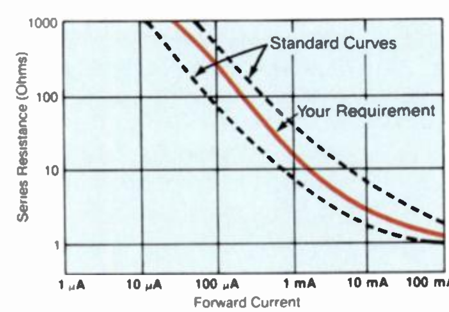

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DSB 6419-55	25	1200	—	800	0.3	100	1.5
DSB 6419-59	40	1500	8.0	1000	0.4	100	2.5
DSB 6419-61	60	1500	—	1300	0.4	100	3.5
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cause of the excessive secrecy desired by most programmers, so the engineer in using them, must rely on the competence of the person writing them and sometimes the assumptions used. As a professional engineer, I prefer to know how a program was developed before paying a lot of money for it.

Perhaps some manufacturer would be interested in developing a combination machine with a calculator and computer keyboard together. It is certainly faster to press the button 1/x than to perform the operation in BASIC.

Yours truly,
Jon GrosJean

"Divider Time Delay"

The "divider time delay" discussed by Mr. Przedpelski in your July/August 1983 issue is an important effect in phase-locked frequency synthesizer design but it is not as simple as is suggested there.

It is true that information cannot be passed from frequency divider to phase detector except at divider output transitions. This happens, at most, twice per reference period and, with some types of phase detectors (phase/frequency, sample-and-hold), only once per reference period. However, this does not necessarily cause a delay equal to the reference period (or to $(N-1) / (F_R N)$ as shown below your Eq. (11)). For example, a sudden phase change at the VCO output could be seen at the phase detector almost immediately if it happened to occur just before the divider's output transition, or it could be delayed as much as a reference period if it occurred slightly later.

What is operating here is a sampling process. If a sine wave of phase modulation should exist at the input to the frequency divider, the width of the narrow pulses from a phase/fre-

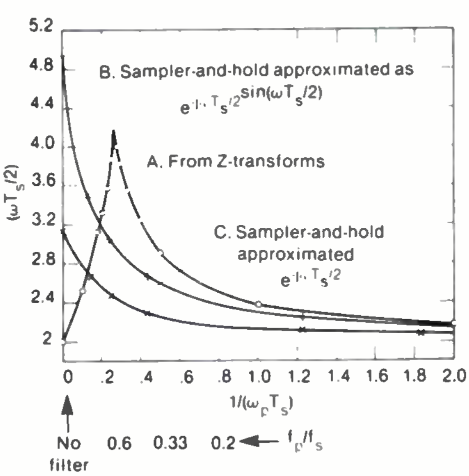
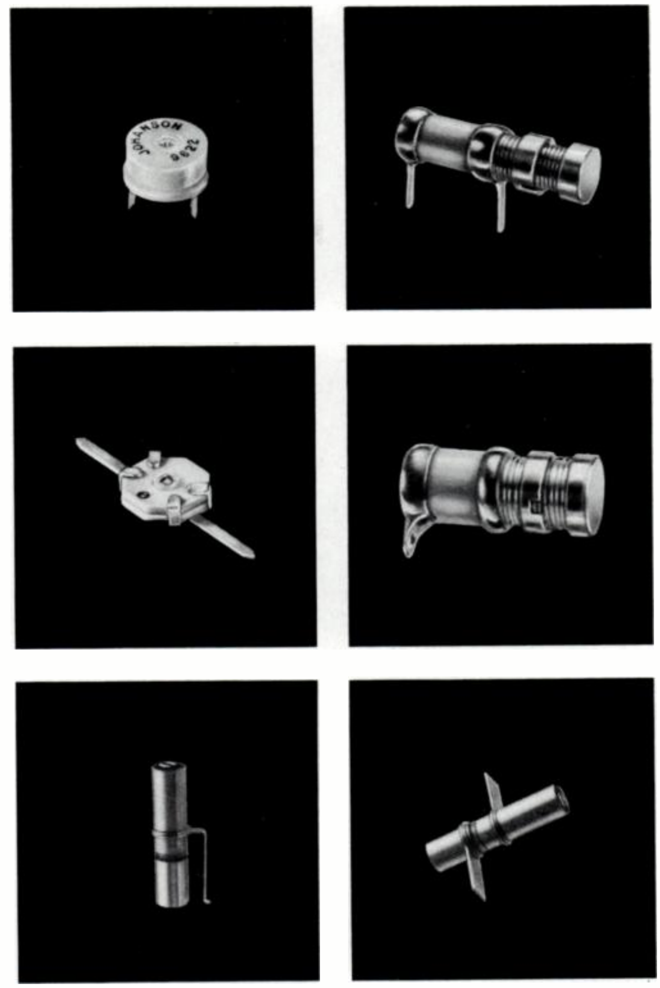


Figure 5.40 Maximum loop gain ω_0 for stable operation versus low-pass corner frequency ω_p, both normalized to sampling period T_s. From **Frequency Synthesis by Phase Lock**, p. 135. Copyright 1981, John Wiley & Sons, Inc. Reprinted by permission.



quency detector (e.g. MC4044) would represent the divided and sampled sine wave and the "fundamental" component of that pulse train would not be delayed relative to the input phase modulation. Of course other newly-created frequencies would also exist at the phase-detector output.[1]

The sampling rate is signal dependent but for small percentage frequency changes it is approximately equal to the reference frequency into the phase detector. Constant-sampling-rate systems can be analyzed exactly using Z-transforms. One approximation sometimes employed where there is a sample-and-hold circuit, such as with a sample-and-hold phase detector, is to ignore all but the lowest (input) frequency generated by the sampling process. The sampled system is then replaced by an approximately equivalent continuouis system. In this case the zero-order hold has a transfer function of

$$e^{-j(\omega T_s/2)} \frac{\sin(\omega T_s/2)}{(\omega T_s/2)}$$

where T_s is the sample period equal to the reciprocal of the reference frequency. A futher simplification retains only the phase term. This gives a time delay of approximately half the value indicated by Mr. Przėdpelski but it is actually associated with the hold circuit rather than the divider. Figure 5.40 (reference 1) shows how these approximations compare to the exact analysis of a simple phase-locked loop with a sample-and-hold phase detector and a loop filter consisting of a single-pole low-pass at ω_p.[2] The verticle scale gives maximum allowed gain ω_o (at $\omega = 1$) for stability. The approximations are seen to be good if the low-pass corner frequency (ω_p) is low compared to the sampling frequency ($1/T_s$).

Other important effects due to this inherent sampling process are the translation of noise frequencies [1,4] and the existance, under some circumstances, of loop oscillations of finite magnitude — that is, they neither grow nor die out.[3]

On a different topic in Mr. Przedpelski's article, the transfer function shown in Fig. 18(b) is for a zero-impedance source. When a charge pump (such as is included in the Motorola MC4044 1C) is used, however, the source is an open circuit at times, usually most of the time, which surely complicates the generation of a transfer function for this filter.

Sincerely,

William F. Egan
Sylvania Systems Group —
Western Division
GTE Communications Products Corp.

1. W.F. Egan, *"Frequency Synthesis by Phase Lock,"* (New York, N.Y.: John Wiley, 1981), pp. 75-81.
2. Ibid, pp. 131-135.
3. Ibid, pp. 209-211.
4. W.F. Egan, "The Effects of Small Contaminating Signals in Nonlinear Elements Used in Frequency Synthesis and Conversion," Proceedings of the IEEE, July 1981, pp. 797-811.

Dear Editor:

I appreciate Dr. Egan's comments. The point of the PLL articles was to provide the designer with some degree of assurance that the circuit he was designing would perform more-or-less as expected. We found that using the standard references this was almost impossible in most instances without deriving the presented formulae, which were not available in literature. It is difficult to put all the information into an article that is not available even in textbooks, but it has been our exper-

ience that, using the material presented, predictable PLL circuits can be built. To comment on Dr. Egan's comments:

- Delay in frequency dividers is seldom mentioned in PLL references. Even Dr. Egan in his book (which I consider as one of the basic reference texts on the subject) talks about divider delay as the propagation delay, which is a function of the logic used and the divider configuration. This is not the delay in question (and usually is small enough, in comparison, to be neglected). Westenkamp seems to be the first to try to reduce this delay to some manageable form. The inclusion of this delay in the overall PLL equation seems to have removed the last discrepancies between paper design and actual hardware in the more complex PLL designs. This delay would be independent of the type of phase detector used.
- Yes, it is, of course, true that equation of Fig. 18(b) assumes a zero source impedance. Any source resistance has to be included in the first R_a resistor. This is common practice and works well with phase detectors such as Motorola MC12040 and MC4044 (without the charge pump — Motorola does not recommend the charge pump for critical applications).

Again, I would like to thank Dr. Egan for taking the time to send his comments.

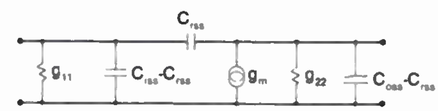
Yours very truly,

Andrzej B. Przedpelski

Diagram Error

Dear Editor,

In the article November/December (1983) Issue "How The Isofet Enhances Stability in Broadband High Gain Amplifiers" some errors were found in Figure 1. More specifically,



Please publish this in the letters or corrections section so that this oversight on my part can be corrected.

I would like to thank Nathan O. Sokal, President of Design Automation, Inc., for informing me of this error.

Larry Leighton

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INFO/CARD 10

Designing Transistor Test Fixtures For The 800 MHz Frequency Spectrum

Test fixtures tend to be the last consideration for most RF power amplifier development programs, yet they are the most valuable tool available for measuring and maintaining device consistency.

By Dan Moline
Motorola Inc.
Semiconductor Products Sector
Phoenix, AZ 85008

Minimum power gain, collector efficiency and broadband performance requirements, though they are always detailed in some form of written specification, are meaningless unless they are demonstrated and controlled by a test fixture. A good test fixture will assure correlation between the customer and vendor while functioning as a trouble shooting tool in the event of radio problems. When alternate sources are pursued for a stage, test fixtures can shorten qualification cycles. But the prevention of gradual shifts in RF performance over the lifetime of a product is the major purpose of a test fixture. Although this article presents techniques for the general case of UHF-800 MHz circuit design, the emphasis is placed specifically on test fixture design for 800 MHz.

Motorola has recognized the importance for good test fixtures and has established general guidelines for their implementation.

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Broadband performance, demonstrating typical characteristics throughout the band. (Ex.: UHF; 450-512 MHz, 800; 800-870 MHz)

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Devices will meet all minimum test requirements at the specified test frequency. UHF: 470 MHz, 800: 870 MHz.

The repeatability, mechanical ruggedness and broadband performance are all very important factors needing consideration in the design of test fixtures. The remainder of this article goes into detail, using the MRF846 as an example.

The schematic representation of the fixture outlined in this article is shown in Figure 1.

C_I and C_O represent the shunt capacitors at the input and output (respectively) which cancel most of the inductive reactance associated with the transistor's input and output impedance. Mini clamped-mica capacitors are used for these components and are physically located beneath the common lead wear blocks. Inductance "L" is introduced by the input (and output) wear blocks. Because of this parasitic inductance, L, trimmer capacitors (C'_I or C'_O) are required to transform the now reactive impedance back to real before launching off into the λ/4 transmission lines.

The transistor's input and output impedance can be represented as a combined series resistor and inductor as shown in Figure 2.

This series combination can be transformed into a parallel equivalent by using the equations shown in Figure 3. The capacitors C_I and C_O are selected by calculating the value necessary to form a parallel resonance with X_p. Since all capacitors have a finite, series lead inductance, the capacitor is actually considered as a simple series resonant circuit. The resulting effect is the capacitance is always higher than the marked value and goes through resonance at some frequency. Mini clamped-mica capacitors are recommended for test fixture design due to the very low parasitic inductance associated with them; which increases the usable range of capacitances. (They are also extremely high "Q"). A typical measured series inductance for clamped-mica capacitors is about 0.5 nH. The equivalent capacitance is calculated by subtracting the series lead inductance from the capacitive reactance, or $X_{c(equiv)} = X_C - X_{L(0.5nH)}$.

Since two capacitors are used in parallel, the total capacitance is derived as shown in Figure 4.

A value of $2X_C$ is used in the example since each capacitor will contribute only ½ to the total capacitance. By setting $X_{C(equiv)}$ equal to the parallel equivalent reactance calculated in Figure 3, the exact capacitor values may be determined.

$$X_p = \frac{2X_c - X_{L(0.5nH)}}{2}$$

$$X_c = \frac{2X_p + X_{L(0.5nH)}}{2}$$

$$(X_c = 1/2\,fC)$$

$$C = \frac{1}{\pi f\,(2X_p + X_{L(0.5nH)})}$$

Introducing an actual example at this time should help in explaining the remaining steps involved in a test fix-

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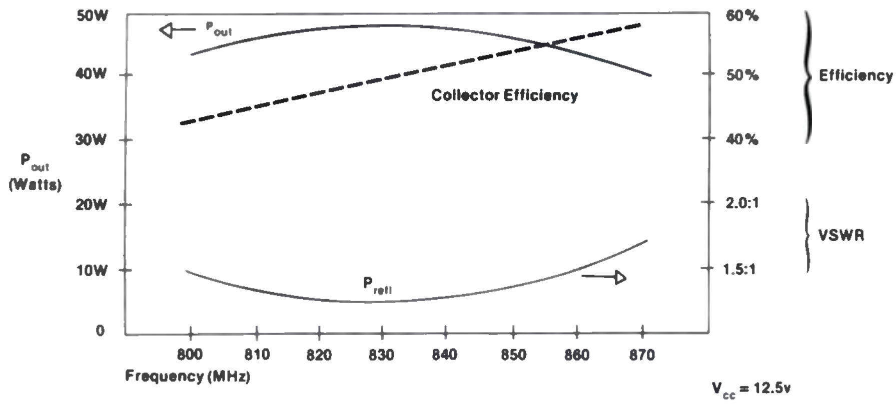

Graph 1.

ture design. The MRF846 is a 40 W, 12.5 V, 800 MHz device whose input and output impedances are:

TABLE 1.

Freq.	Zin	Zout
800 MHz	1.1 − j4.8	1.20 + j2.4
836 MHz	1.0 − j4.9	1.15 + j2.5
870 MHz	1.0 + j5.0	1.10 + j2.7
900 MHz	0.9 + j5.1	1.10 + j2.8

Since X_p will vary as a function of frequency, C_I and C_O need only be calculated for one point within the frequency band. Typically, the input response of an RF power tansistor is optimized about the center of the band. Hence, the input R_p and X_p are generally calculated at this frequency $[(f_H + f_L)/2]$

The output response is different. If C_O were selected for a resonance to occur with X_p, at band-center, an unacceptable performance roll-off would be seen at the upper end of the frequency band. Overall performance is best when C_O is calculated at a frequency within 20% of the upper end of the band. Since device gain increases as frequency decreases, the performance at lower frequencies is generally no problem.

Using the MRF846 as an example, input and output capacitor values may be determined as follows:

INPUT: Freq. = 836 MHz

$Zin = 1 + j4.9 \qquad Q = 4.9/1 = 4.9$

$X_p = 4.9(1 + 1/(4.9)2) = 5.1$ ohms

$X_{L(0.5nH)} = 2\pi(836 \times 10^6)(.5X10\text{-}9)$
$= 2.63$ ohms

$C = 1/[\pi(836 \times 10^6)(2X5.1 + 2.63)]$
$= 29.7$ pF

2-15 pF Capacitors would be the best choice.

OUTPUT: Freq. = 870 MHz

$Z_0 = 1.1 + j2.7 \qquad Q = 2.7/1.1 = 2.45$

$X_p = 2.7(1 + 1/(2.45)2) = 3.15$ ohms

$X_{L(0.5nH)} = 2\pi(870 \times 10^6)(.5X10\text{-}9) = 2.7$ ohms

$C = 1/[\pi(870 \times 10^6)(2X3.15 + 2.7)]$
$= 40.7$ pF

2-20 pF Capacitors would be the best choice.
(20 pF Capacitors were not available, so an 18 pF & a 24 pF capacitor were chosen instead. The total C = 42 pF).

Though the MRF846 test fixture used at Motorola does use these capacitor values, the above calculations may act only as a good starting point. Imperical measurements and more precise impedance measurements for a given applications may

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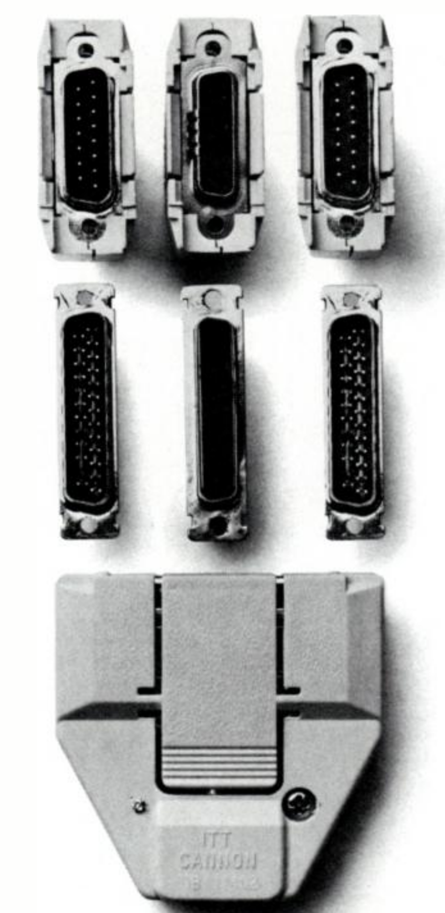

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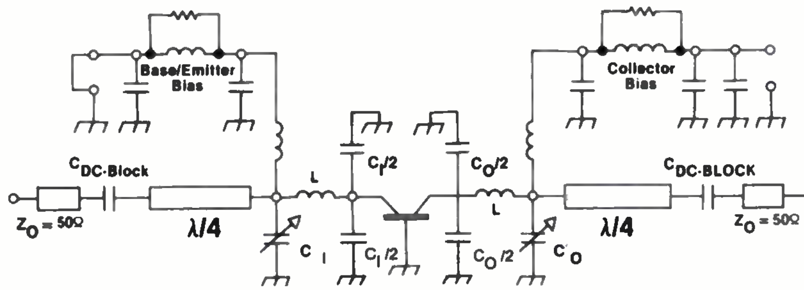


Figure 1

result in minor deviations from these values.

Assuming no additional circuit parasitics had to be accounted for, the quarter wave transmission line sections could now be determined. the input (and output) fixture wear blocks do, however, contribute additional series lead inductance to the impedances. These inductances are counteracted by the trim capacitors C'_I and C'_O. The wear block inductance could be calculated and then used to determine the proper capacitance values. However, since there are other, less obvious frequency and grounding effects which may influence the impedance transformation, it is a more practical (and generally a more accurate) procedure to measure the impedance which will be transformed by the transmission line to 50 ohms.

The capacitors C_I and C_O should be mounted into the test fixture and a known characteristic impedance transmission line soldered into place as shown in Figure 5.

Triple stub tuners are used on the input and output to tune for maximum output power and minimum reflected power at various frequencies throughout the band. Band edges and band center are generally adequate for a good circuit design. Due to higher impedance levels produced by adding C_I and C_O, [Zin] and [Zout] are measured instead of the real transistor impedances, Zin and Zout. Also, by measuring impedances in the actual applications fixture, the design can be optimized for that particular fixture. Perhaps a maximum gain tuning point is not desired. Obtaining impedances for an efficiency/gain compromise may

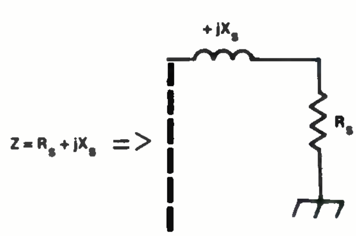


Figure 2

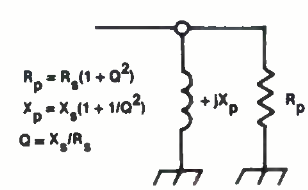


Figure 3

$$X_{C(equiv)} = \frac{2X_C \cdot X_{L(.5nH)}}{2}$$

Xc (equiv) = >

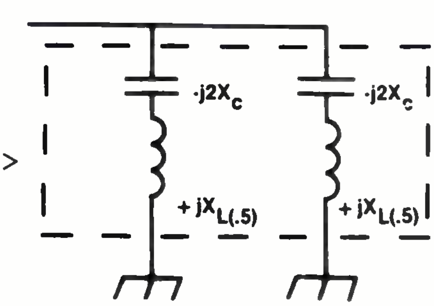
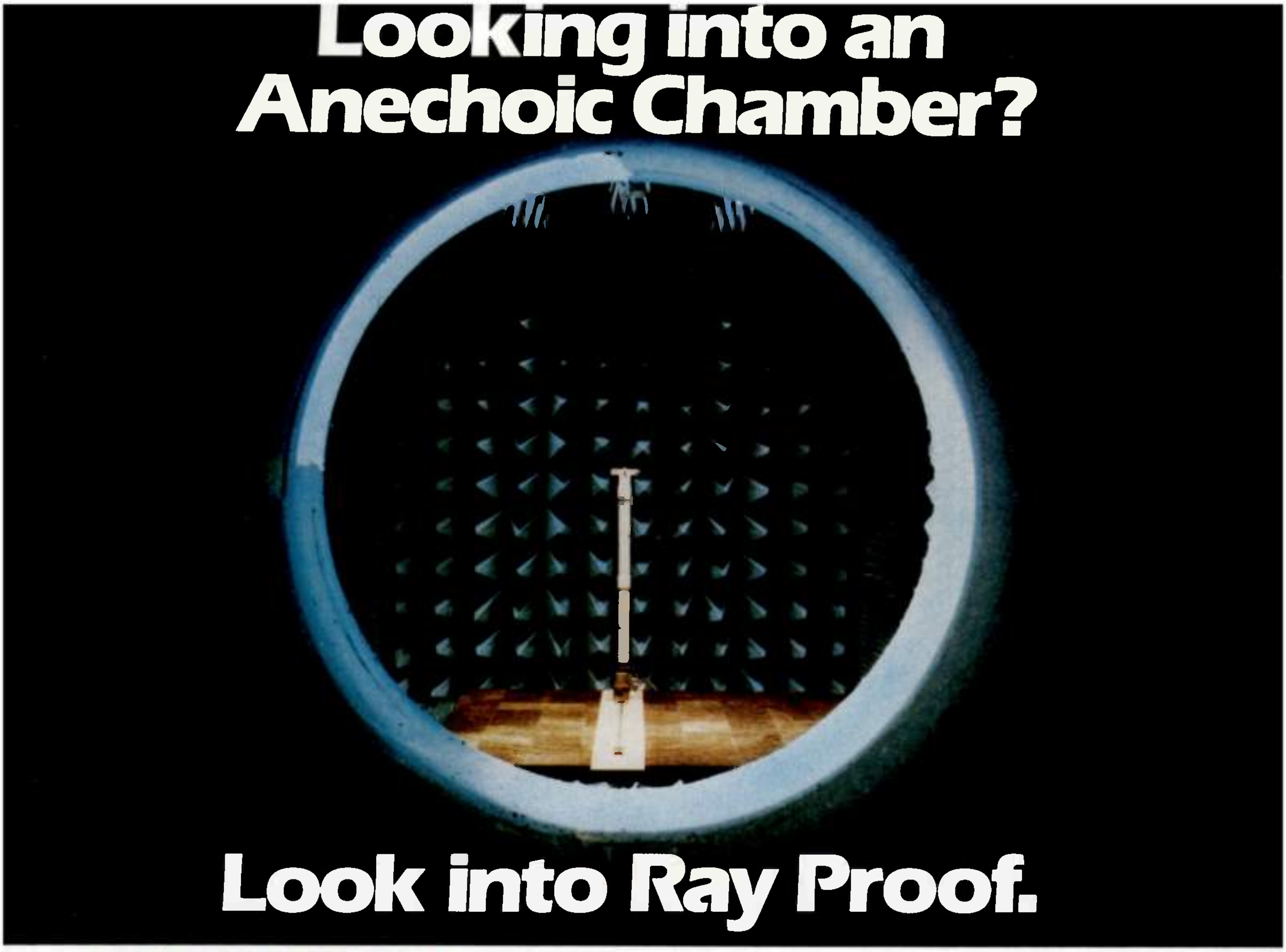
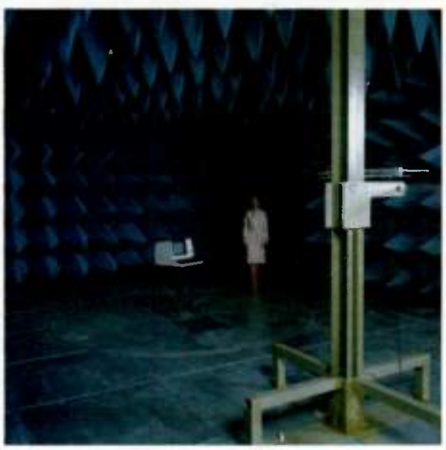


Figure 4





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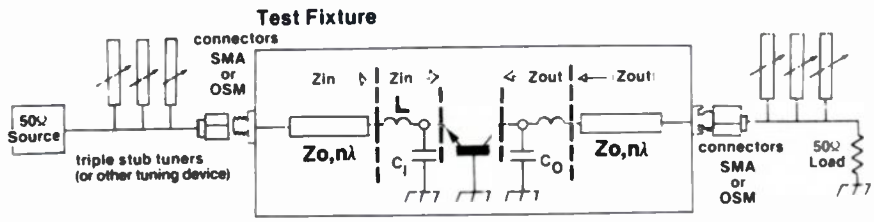

Figure 5

be more desirable. If this is the case, an impedance table for the appropriate conditions may be obtained. It is then for these impedances that C_I, C_O and Z_O will be calculated.

The procedure for obtaining the impedances is simple and requires a vector voltmeter (VVM) or a network analyzer. Both are used at Motorola, but a vector voltmeter is less expensive and if used with a high directivity directional coupler, (>40 db), is very accurate. The set-up is constructed as shown in Figure 5. With the frequency set, stub tuners are adjusted for the desired performance. Again, using the MRF846 as an example, the following numbers were measured:

Pin = 12.0 watts, Vcc = 12.5 volts

806 MHz	838 MHz	870 MHz
P_{out} = 50.0W	P_{out} = 48.3W	P_{out} = 44W
Eff. = 53.3%	Eff. = 55.2%	Eff. = 58%
$P_{refl.}$ = 0W	$P_{refl.}$ = 0W	$P_{refl.}$ = 0W

The output stub tuners were adjusted for maximum gain at each frequency and the input stub tuners were adjusted for 0 watts reflected power. After each measurement, the impedance presented to the fixture by the triple stub tuner and load (or source) combination is measured by the vector voltmeter. The impedance is then translated by the transmission line used in the test fixture to obtain [Zin] and [Zout]. In the above example, a 26 ohm, .309λ (@836 MHz) transmission line was arbitrarily chosen to be in the MRF846 measurements. By using the equation: $Z\angle\theta = R_O[(1+\Gamma\angle\theta)]/(1-\Gamma\angle\theta)]$. or various computer or calculator programs, the transformation is easily calculated. The most important part of the whole procedure is obtaining an accurate measurement from the stub tuners. Prior to making any measurements, the vector voltmeter must be referenced to a short (180° on a Smith Chart). As a means of accounting for the errors introduced by the connectors at the fixture's input and output, that same connector is used for a referencing short as shown in Figure 6.

The measurement reference plane is now the edge of the connector used on the test fixture, which is also the beginning of the transmission line. Assuming the same reference plane is maintained during the measurements, an accurate impedance value will be produced. A good technique for maintaining the appropriate reference plane is accomplished by creating a new connector to measure the triple stub tuners. Two connectors are attached as shown in Figure 7.

The triple stub tuner, load combination may now be measured with an adequate degree of accuracy, as demonstrated in Figure 8.

Repeat the process for the input stub tuner combination. Two numbers are obtained for each frequency from which [Zin] and [Zout] can be calculated from, as shown in the MRF846 example below:

Freq.		Measured $\Gamma\angle\theta$	$\Gamma\angle\theta$ converted to impedance	Impedance Transformed over 26Ω line	
806 MHz	Input	0.35∠155	24.97 + j8.42	20.72-j5.64	[Zin]*
	Output	0.37∠144	24.86 + j12.53	17,72-j6.66	[Zout]*
838 MHz	Input	0.26∠166	29.78 + j3.98	21.35 + j0	[Zin]*
	Output	0.22∠154	33.30 + j6.58	18.68 + j.74	[Zout]*
870 MHz	Input	0.14∠-169	38.25-j1.99	20.21 + j6.92	[Zin]*
	Output	0.07∠-158	44.10-j2.24	17.90 + j8.3	[Zout]*

Table 3

Note: [Zout]* is conjugate of [Zout]   [Zin]* is conjugate of [Zin]

The new impedances can be obtained by using a Smith Chart or using the equation $Z/\theta = R_O[(1+\Gamma/\theta)/(1-\Gamma/\theta)]$. These impedances (shown in the last column of Table 3) are the impedances around which the test fixture will be optimized. Once again, it is convenient to convert these numbers into parallel equivalents. By doing so, the values of C'_I and C'_O become more obvious. Table 4 shows this process.

Series Impedance [Zin] & [Zout]	R_p	X_p	Capacitance Required	
20.72 + j5.64	22.26	j81.8	2.42 pF	C'_I
17.72 + j6.66	20.2	j53.8	3.67 pF	C'_O
21.35 + j0	21.35	∞	0 pF	C'_I
18.68-j.74	18.7	j472	.40 pF	C'_O
20.21-j6.92	21.6	-j68.3	-2.68 pF	C'_I
17.90-j8.3	21.75	-j46.9	-3.90 pF	C'_O

Table 4

From Table 4, notice the calculated values of C_I and C_O come close to giving the desired frequency response. C'_I is zero at the band center, indicating the capacitors selected for the input is optimum. The values for C'_O produce a slight skew in performance toward the high end of the band. Capacitor values for the output could be reduced slightly, but they will remain the same until final fixture performance is determined.

Since C'_I and C'_O are very small capacitor values, little or no capacitance is actually needed for C'_I and C'_O. However, to allow minor tuning adjustments, a small trimmer capacitor is included at the wear block/transmission line interface.

The final calculation which needs to be performed is that of finding the optimum characteristic impedance for the transmission line. The recommended approach for doing this is to use a computer optimization program which will iterate any number of variables for a desired frequency response. The variables available to be optimized at this point are Z_O. C'_I and C'_O and even nλ (transmission line length) Z_O, C'_I and C'_O are the very minimum variables.

In the example of the MRF846, where input Rp varies from 22.3 ohms to 21.6 ohms over the frequency band, a close approximation can be had by using a mean value of 21.9 ohms. This results in a Z_O of $\sqrt{50 \times 21.9} = 33$ ohms The output Rp starts at 20.2 ohms, dips to 19.7 ohms and goes back up to 21.75 ohms. Using the same method as before, Z_O is calculated as $\sqrt{50 \times 20.2} = 31.7$ ohms where 20.2 ohms is the mean value of 18.7 and 21.75. Using

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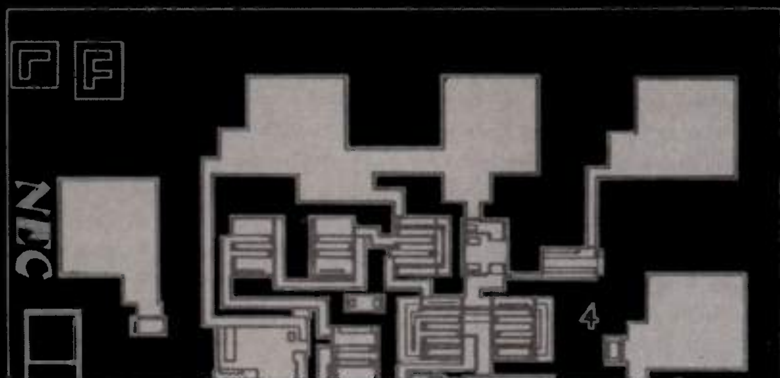
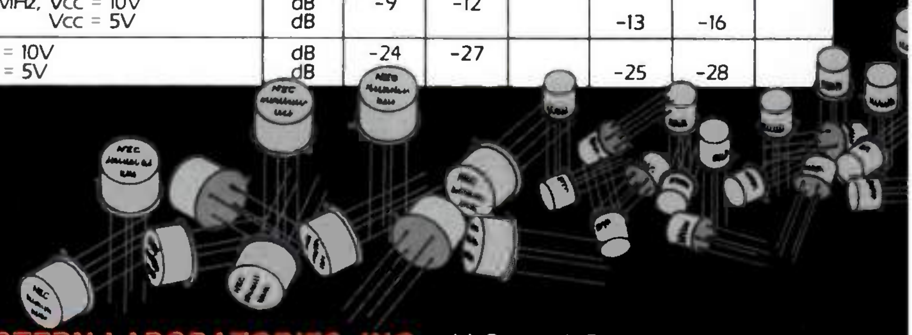


	MM PART NUMBER Package Code		MM76515 15 (TO-33)			MM76612 12(TO-72)		
SYMBOLS	PARAMETERS AND CONDITIONS	UNITS	MIN	TYP	MAX	MIN	TYP	MAX
Icc	Operating Current at Vcc = 10V Vcc = 5V	mA mA	38	43	48	15	20	25
|S21|	Power Gain at f = 500 MHz, Vcc = 10V Vcc = 5V	dB dB	17	19	21	16	18	20
NF	Noise Figure at f = 500 MHz, Vcc = 10V Vcc = 5V	dB dB		5.5	6.5		5.5	6.5
BW	Bandwidth at Vcc = 10V, (3 dB BW) Vcc = 5V, (3 dB BW)	MHz MHz	950	1100		1100	1300	
P1dB	Output Power (1dBc) at f = 500 MHz, Vcc = 10V Vcc = 5V	dBm dBm	9	10.5		3	5	
IM3	3rd Order Intermodulation at f1 = 500 MHz, f2 = 510 MHz f = 2f1 − f2, Vcc = 10V, Pout = 0dBm Vcc = 5V, Pout = −7dBm	dB dB		-40			-40	
|S11|	Input Return Loss at f = 500 MHz, Vcc = 10V Vcc = 5V	dB dB	-15	-18		-20	-23	
|S22|	Output Return Loss at f = 500 MHz, Vcc = 10V Vcc = 5V	dB dB	-9	-12		-13	-16	
|S12|	Isolation at f = 500 MHz, Vcc = 10V Vcc = 5V	dB dB	-24	-27		-25	-28	



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In Europe Contact: NEC Electronics (Europe)GmbH • Oberrather Strasse 4 • 4000 Düsseldorf 30 • West Germany a computer optimization program, a 32 ohm, quarter wave transmission line is optimum for the input and a 30 ohm quarter wave line is optimum for the output. These are the values used in the MRF846 test fixture.

After constructing the MRF846 test fixture and tuning the small trimmer capacitors for best overall gain and input reflected response, the average data shown in Graph 1 was obtained.

The goal was to demonstrate 12/40 W across the band with better than 2.0:1 input VSWR and better than 45% collector efficiency. Further optimization could be done by performing impedance measurements on additional transistors or characterizing the test fixture more accurately. However, the above performance is very satisfactory to the required performance. The best compromise for a second pass fixture would be to tradeoff gain at 806 and 838 MHz for efficiency, and redesign input and output matching networks for the new impedance tables. This, of course, is only one of the many procedures which may be followed in developing an 800 MHz test fixture.

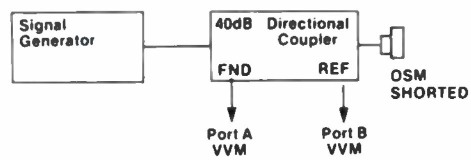
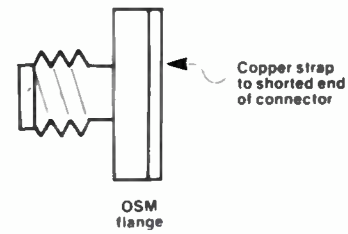


FIG. 6



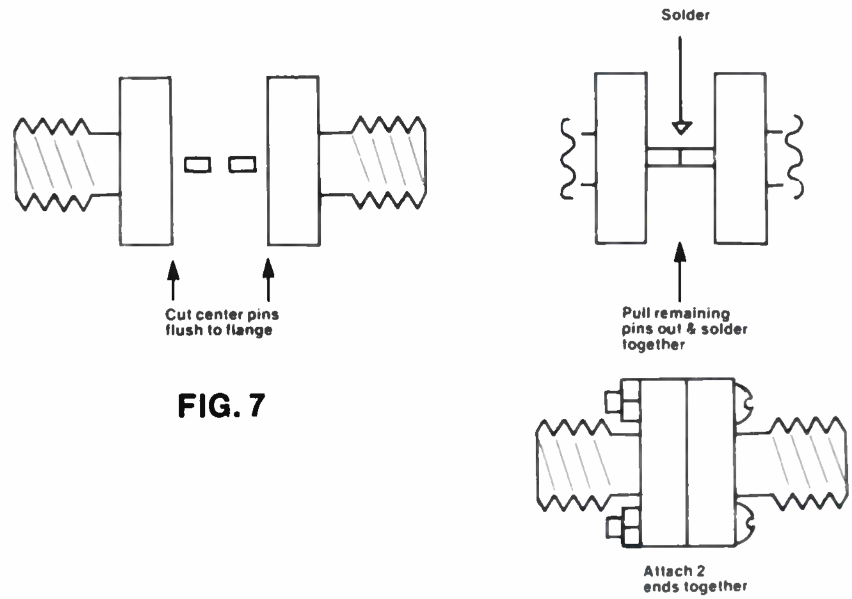


FIG. 7

INFO/CARD 18

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INFO/CARD 19

Some Questions You Should Ask Your EMI/RFI Test Equipment Supplier

How long has your company been in the EMI/RFI Testing Equipment business?

The Electro-Mechanics Company has been designing and manufacturing EMI/RFI Test Accessories for over 25 years.

We began with Magnetic Field Instruments and Radio Frequency Antennas for the military. Since the mid-sixties we have been working closely with the military and industry standard setting groups, and EMCO now has one of the broadest lines of Test Antennas and EMI/RFI Test Accessories in the World.

What exactly is your major product line?

EMCO's primary business is Test Antennas for use in Emissions and Immunity (Susceptibility) Testing as required for MIL Standard, FCC, VDE, and CISPR Test Procedures.

Typical Military Antennas for Radiated Immunity (RS) and Emissions (RE) Testing cover the frequency ranges from **30 Hz to 18 GHz**, and are noted in MIL STD 462, Notice 3, Table 1. EMCO currently manufactúres **Magnetic Field Loops**, the 41" **Rod Antenna, Parallel Strip Line**, both **Biconical Antennas**, the **Conical Log Spirals** and the **Double Ridged Guide Antennas** shown on this table.

Antennas which are currently acceptable for use in FCC Volume II, Part 15 Emissions Testing include, **Adjustable Element Dipole Sets**, Broadband **Biconical Antennas** and Broadband **Log Periodic Antennas**. EMCO manufactures all of these separately or can include them as part of an FCC "Class A" and "Class B" **Antenna Test System**.

What differentiates your antennas from your competitors?

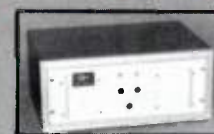
One major difference is Calibration. Each Antenna is calibrated using NBS Traceable Testing Equipment, on our own FCC open field test site. Calibration data includes Antenna Factor, Numeric Power Gain, and dBi Gain for each individual Antenna. For Immunity Testing Antennas we include Field Strength measurements in Volts Per Meter, and Radiation Patterns where applicable.

Another difference is Design and Construction. Each Antenna is designed to be durable and long-lasting, yet functional in varied applications, such as in Anechoic Chambers or Outside Test Sites. Antennas and accessories are machined and constructed "in-house" for Optimum Quality Control.

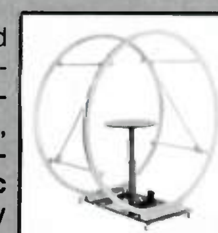
One last difference and maybe the most important, is EMCO's continual Product improvement thru Research and Development. For example, our **Dipole** and **Biconical Balun** design is much improved from the old DM-105 and military designs . . . and we are continually researching and redesigning to make EMI/RFI Testing simpler and more accurate.

What other Test Equipment and Accessories do you offer?

EMCO adds efficiency to EMI Testing with an **Antenna Positioning Tower** (1-6 meters) and an **Equipment Testing Turntable**. Both are suitable for outside or indoor use, come with a standard Digital Readout Controller and are available with **IEEE-488 Bus** Option.



For Conducted Emissions Testing, EMCO manufactures **Line Impedance Stabilization Networks** to satisfy FCC and VDE requirements. Our unique design allows production of as many as 4 separate lines (three phase) in one unit.



Other Related Equipment include: **Signal Rejection Networks, Acceptance Networks, Magnetic Field Intensity Meters, Magnetometers** and **Helmholtz Coil Systems.**

Why should my company buy your EMI/RFI Test Equipment?

The Electro-Mechanics Company is more than just another manufacturer. We realize that in order to grow and help improve EMI/RFI Testing we must constantly forge ahead . . . not live in the past.

As the FCC moves toward better and more Standardized Test Procedures, EMCO is staying close to ANSI (American National Standards Institute), NBS (National Bureau of Standards) and other standards groups so we can keep improving our equipment. Involvement with current and future industry needs also helps us plan for design of new equipment . . . an ongoing process at EMCO.

EMCO is committed to offering Technical Assistance, as well as Test Accessories, to help solve EMI Testing Problems. Part of that Technical Assistance is advice on purchasing only the equipment needed, not kits or systems with unnecessary items. We can also advise on various manufacturers of other complimentary test equipment.

If you have more questions and are looking for Helpful Answers, Call us at (512) 835-4684.

The Electro-Mechanics Company

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INFO/CARD 20

High Performance VSWR Measurements

The Return Loss Bridge is a useful device for making low power precision VSWR measurements. This article describes the operation and use of this instrument.

By James S. Beck
Eagle Consultants
Fallbrook, CA 92028

This article describes the operation and use of the Return Loss Bridge. This includes a basic definition of terms, outlines of test and other background information. Although there are many methods available to perform VSWR measurements, ranging from sliding lamps down a transmission line to highly precise slotted lines, only directional bridges will be addressed in this article.

Background

What is *VSWR, rho,* and *return loss?*

When reading specifications describing various equipment one often see references to VSWR, or possibly rho or return loss. These are all expressions for the same quantity and if one is known the others can be determined by simple mathematics.

These quantities are defined as follows:

VSWR: This is the maximum amplitude on a transmission line to the minimum amplitude on the transmission line; this is expressed mathematically as:

$$\text{VSWR} = \frac{\text{V(MAX)}}{\text{V(MIN)}}$$

ρ : Is the Greek letter *"rho"* and expresses reflection coefficient; this is described mathematically as:

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

It can be seen when VSWR is infinite $\rho = 1$ and when the VSWR = 1 then $\rho = 0$. By multiplying ρ times 100 and squaring the percentage of power reflected into the source is found.

Return Loss: This is simply the logarithmic expression of rho. This is defined as return loss (dB) = $-20 \log_{10} \rho$. This quantity is the amount of power returned to the source relative to the amount of power delivered.

Return Loss Conversion Chart

Return Loss	Rho	SWR
1.0	0.89	17.4
3.0	0.71	3.0
5.0	.56	3.6
10.0	.32	1.9
15.0	.18	1.4
20.0	.10	1.2
25.0	.056	1.12
30.0	.032	1.07
35.0	.018	1.04
40.0	.010	1.02
50.0	.003	1.006
60.0	.001	1.002

Table 1

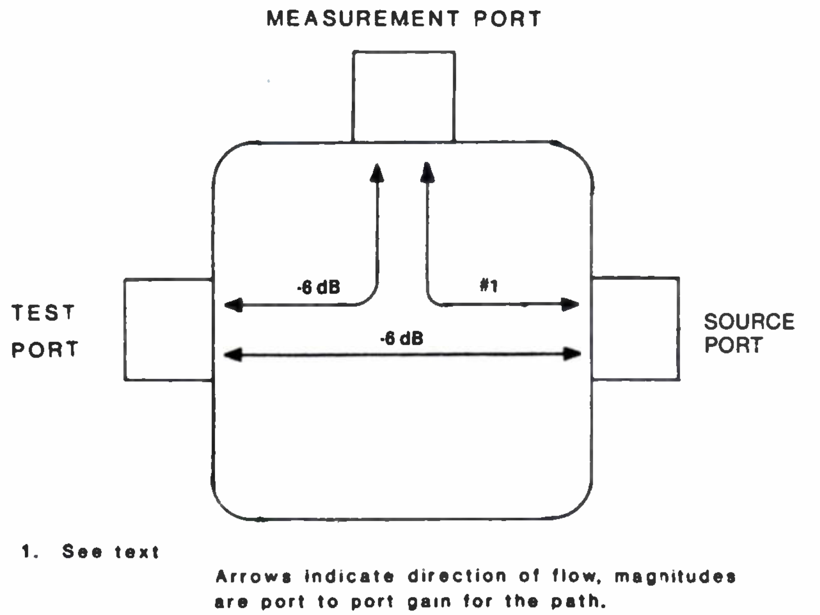
Measurements

The measurement equipment will determine the type of data that will be presented. Use of a VSWR meter will provide a scale that presents the data in terms of VSWR. If directional couplers are used then the absolute reflected power will be known and calculations will put this data into any desired form. A directional bridge will present either rho or return loss depending on the ancillary equipment. It should be noted that directional bridges appear under many names which can be confusing. A directional bridge may be referred to as a rho bridge, return loss bridge, directional bridge or impedance bridge. Generally speaking a bridge that has an internal detector is called a rho bridge while bridges with an RF output use the other terms.

Directional Bridges

Directional Bridges are available in all form and fashion, frequencies range from audio to microwave, some contain internal references while others do not. Bridges may contain internal detectors at the MEASUREMENT PORT or the port may be of the RF variety; finally the price range of these bridges is from $100 to $2000.

Figure 1 is illustrative of a typical return loss bridge. The arrows and lines indicate the direction and gain of the power flows through the bridge. For VSWR measurements the source (stimulus or generator) is connected



RETURN LOSS BRIDGE PORT CONFIGURATION

Figure 1

TM8 autoranging millivoltmeter

The Texscan TM8 is an autoranging, true r.m.s. millivoltmeter with a specified operating range of 10kHz to 1GHz with useful indications to 1500MHz. It measures r.f. voltages from 1mV to 3V full scale in eight ranges and may be operated automatically (in the autoranging mode), manually or remotely. The maximum voltage measured can be increased to 300V r.m.s. using the 100:1 divider provided and readings on any range can be held for at least three minutes by momentary depression of the probe switch.

Measurements are true r.m.s. but average responding can also be selected with a front panel switch. The meter reads 100μV-3V and also features a corresponding dB scale from −60 to +23dBm. A single logarithmic range is also provided and covers four decades scaled −30 to +10dBm. Fast or slow meter response may be selected to permit the observation or removal of rapid signal amplitude changes.

Remote programming is a standard feature and the socket is located on the rear panel. An output to drive a pen recorder is also provided from this socket and this can be used with an interface module, OB2, to connect to IEEE-488 bus systems.

The TM8 is supplied complete with probe (integral with input lead), probe to BNC adapter, 'T' connector, and a 100:1 high impedance divider.

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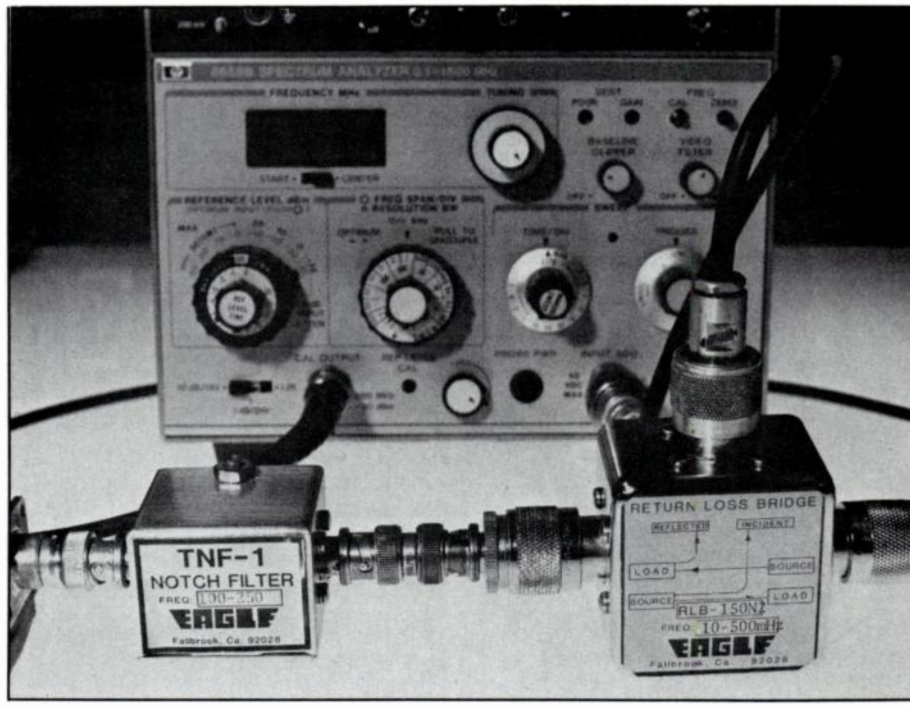

to the right hand port, SOURCE PORT, while the device to be measured (load) is connected to the left hand port, TEST PORT. When the power is applied it flows from the right to left hand port losing 6 dB through the internal resistors in the bridge. If the left hand port is terminated perfectly no power will be reflected from the left hand port. Since no power is reflected there will be no power appearing at the MEASUREMENT PORT, at least theoretically.

Some power will be present at the MEASUREMENT PORT, however, this is due to imperfections in the bridge. This is illustrated by the path from the right hand port, SOURCE PORT, to the top port, MEASUREMENT PORT. The gain of this path will be -52 dB if the bridge has a directivity of -40 dB. The directivity of a bridge is measured by terminating it in a perfect, or near perfect, load and measuring the residual power at the MEASUREMENT PORT. The residual power is then referenced to the amount of power appearing with the TEST PORT terminated in an open or short; the result is converted to dB and becomes the directivity of the bridge. This paramenter is very difficult to measure because of the difficulty in finding "perfect" terminations.

Above 500 MHz or so the connector VSWR becomes significant in precision bridges; thus even if a perfect load is presented to the connector interface there is the possibility that the bridge itself is seeing reflections from its own connector. Bridges operating above 500 MHz with specified directivities of 40 dB or better should use an external reference termination so that the interface connectors at the TEST PORT and REFERENCE PORT are identical. If this is the case then the connector VSWR will factor out.

If the TEST PORT is shorted or opened then all of the power is reflected back to the MEASUREMENT PORT again losing 6 dB due to internal losses in the bridge. Because of the internal losses the infinite VSWR (open or short) will be 12 dB below the amount of power that is present at the SOURCE PORT.

Power may also be applied to the MEASUREMENT PORT. If this is the case it will be transferred to the TEST PORT again losing 6 dB. Thus if a generator is connected to the MEASUREMENT PORT and another generator is connected to the SOURCE PORT the signals from these generators will be combined by the bridge. This combined power will appear at the TEST PORT. If the TEST PORT is terminated in a reasonable load no power will be reflected back from the TEST PORT; thus the generators will be well isolated from each other. This technique is useful in making intermodulation distortion tests and other measurements requiring two generators.

It should be remembered that the losses through the paths in the bridge are theoretical 6 dB. In actual practice these losses will be somewhat greater than 6 dB. In a good bridge 7 to 7.5 dB is not unreasonable at higher frequencies.

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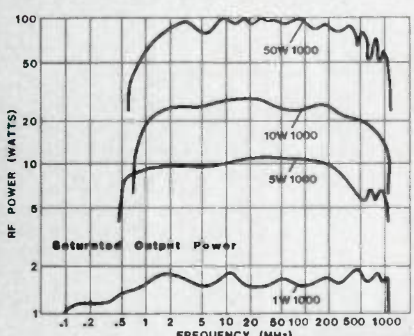

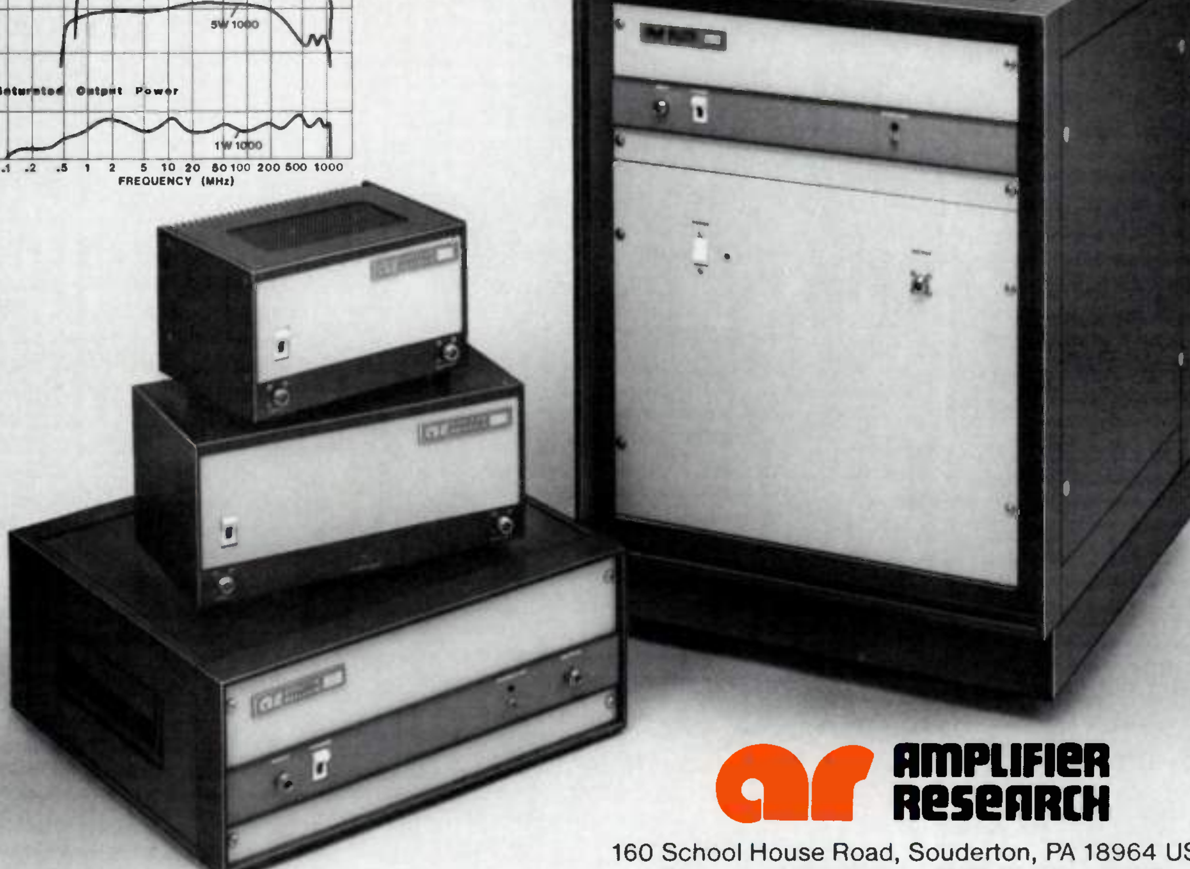

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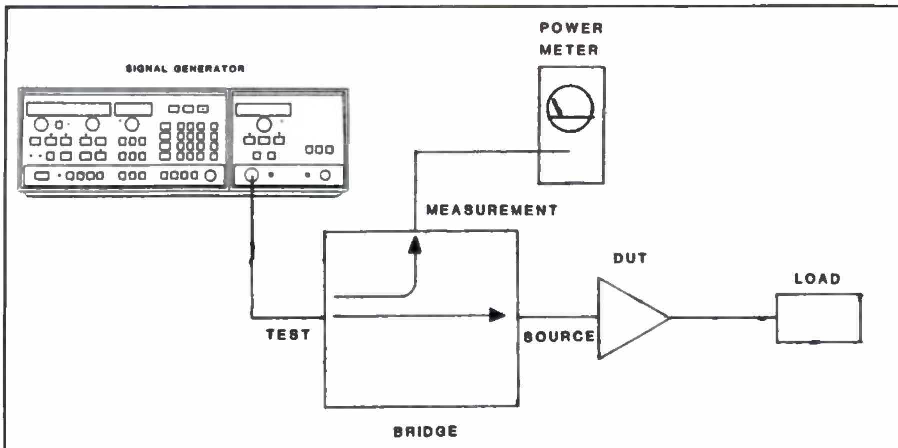

FIGURE 2

Bridges that provide an RF signal at the MEASUREMENT PORT tend to be more expensive because of the increased complexity of manufacture. These bridges are much more versatile. By using a spectrum analyzer several signals can be measured at one time. With an internal diode detector, this would be impossible and the signals would have to be removed in order to get an accurate measurement. With an RF MEASUREMENT PORT, phase information is present which can be very useful. Also, power can be fed into the MEASUREMENT PORT thus allowing the bridge to be used for power combining purposes.

Depending on the application an external or internal reference might be desirable. In most applications an internal reference is preferred because it simplifies the measurement and the requirement to maintain precision terminations. High precision measurements require the connector VSWR to be factored out by using identical connectors on the reference and test ports. On bridges equipped with the "F" type connector this is a must to obtain any reasonable reading at all.

Applications

In addition to the standard function of making return loss measurements, return loss bridges may also be used for other tests that are very common in the RF laboratory. The following uses are just some of the many applications of an instrument equipped with an RF MEASUREMENT PORT.

Drive Power Measurement

In this circuit (Figure 2.) the power meter will read the amount of power being delivered to the DUT, assuming the DUT has the same impedance as the bridge. In this case the power rating of the bridge will determine the amount of drive power that can be applied to the input port. Remember that since the power is split there will be a 6 dB power reduction. Bridges are available with up to a five watt power rating which would allow 1.25 watts drive to the DUT.

Power Combining for TWO-TONE Tests.

In this case (Figure 3.) the output to the TEST PORT will be the combination of signals produced by the two generators. The generator port to port isolation will be approximately the same as the specified directivity of the bridge at the two test frequencies. This feature will aid in preventing intermodulation distortion because the signal generators will be well isolated from each other. The above test set can be used for measuring the IM performance of a DUT. If intermods are present, set the spectrum analyzer to 10 dB of RF attenuation to check that the intermods are not coming from the spectrum analyzer.

Amplifier Input Return Loss

A signal generator (Figure 4.) is connected to the SOURCE PORT of the return loss bridge and the amplifier (DUT) is connected to the TEST PORT. A power meter connected to the MEASUREMENT PORT will now indicate the amount of power reflected by the DUT. Be sure to add six dB to the power meter's reading because of the six dB loss in the arm between the TEST and MEASUREMENT port.

There is also a power loss of six dB between the SOURCE and TEST port; this means that the amount of drive available to the DUT will be reduced by six dB. The maximum amount of drive that can be applied will depend on the power rating of the bridge. A one watt rated bridge will deliver .25 watts while a bridged rated at 5 watts can safely deliver

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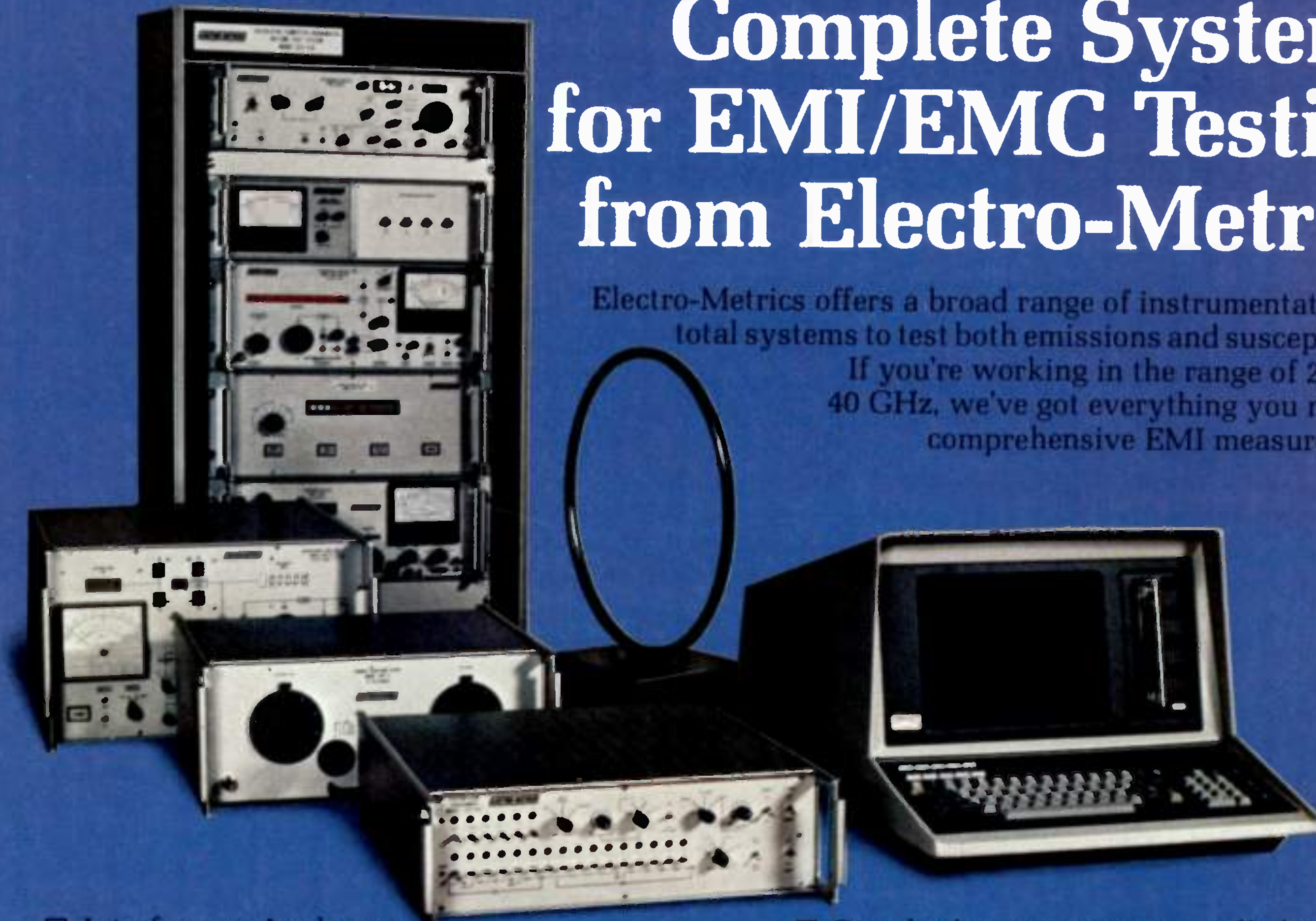

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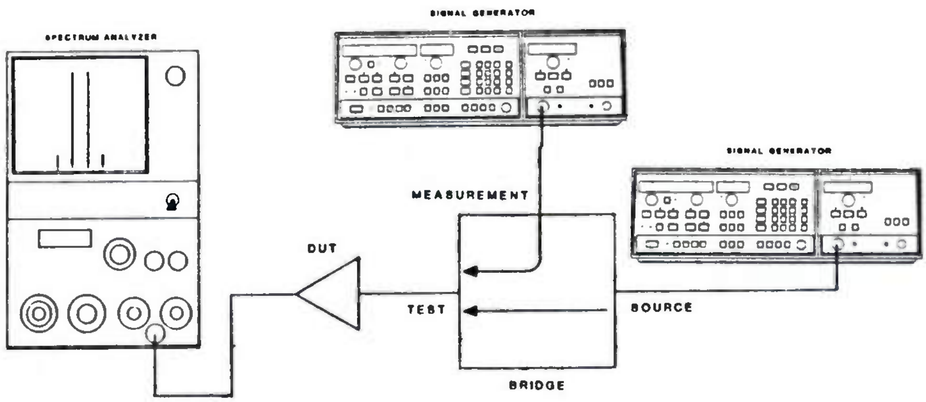


Figure 3

1.25 watts. If more than 1.25 watts is required a different technique will have to be utilized.

Amplifier Output Return Loss

In this test (Figure 5) the amplifier (DUT) is connected to the TEST PORT while an attenuator is connected to the SOURCE PORT. A signal generator is then connected to the other side of the attenuator. A second attenuator is connected to the MEASUREMENT port of the return loss bridge. The attenuators are to protect the spectrum analyzer and signal generator from damage.

The test procedure is as follows:
1. Adjust the signal generator (#2) to the amplifier carrier frequency plus or minus the narrowest separation required by the spectrum analyzer to discriminate between the two signals.
2. Apply enough power from generator 2 so that the power level at SOURCE PORT is approximately 10 dB below output level of amplifier under test.
3. Apply drive to amplifier and adjust for maximum or test output power. *CAUTION! Bridges can handle no more than rated power!!*

Antenna System Return Loss

Antenna measurements can be readily made using return loss bridges. Before covering the actual measurements it will be beneficial to discuss some of the considerations involving real world antenna measurements. Antenna measurements present a unique and sometimes hazardous situation due to congestion and colocation of equipments. Antennas that are connected to high power transmitters may be located within several hundred feet of each other. Thus, there may be considerable energy present at the terminals of the antenna to be tested.

Before commencing any measurement, study the location to determine if any other system could radiate enough power into the antenna under test to cause equipment damage. The chart below will aid in making this determination:

Power Level	Minimum Safe Distance	Power Received
1000 watts	100 feet	.24 watt
10000 watts	500 feet	.10 watt
100000 watts	1000 feet	.24 watt

The above chart is based on both stations using unity gain antennas on an operating frequency of 50 MHz. If frequency is lower gain will be higher and vice versa.

If there is some doubt as to whether the measurement can safely be made or not the exact power can be calculated by the following formula:

$$P_r = P_t G_r G_t \lambda^2/(4\lambda R)^2$$

Where:
λ = wavelength in meters
G_r = gain of receiving ant.
G_t = gain of transmit ant.
P_t = transmit power
R = distance in meters
P_r = received power

After making the calculation if there is still any doubt than connect a power meter such as a Bird thru-line to the antenna system; and measure the power level present. Remember if there is more than one transmitter, the power levels are cumulative.

CAUTION: Insure that the colocated transmitters are on while making antenna power measurement.

As long as the incoming power level is below the rated power level of the bridge the measurement can be safely made. A bridge rated at five watts can normally be used with no problem. The Figure 6 illustrates a typical measurement system for antennas:

The power meter is indicating 1.25 watts at the MEASUREMENT PORT of the return loss bridge; and the signal generator isn't even turned on yet. The power meter is reading the power from the FM station which is being received by antenna at KTV. It was anticipated that the generator would supply a signal of +12 dBm and the directivity of the antenna would be 35 dB. Since 12 dB is lost through the arms of the bridge, the signal returned would be -35 dBm. This is 66 dB below the signal being received from KFM which is +31 dBm. No way can the measurement be made while KFM is on.

If the return loss bridge has a DC MEASUREMENT PORT you are out of luck. Your reflected signal and KFM's are combined by the detection diode; internal in the return loss bridge. The only hope is to shut KFM off while the measurements are being made. If the return loss bridge has an RF MEASUREMENT PORT then simply replace the power meter with a spectrum analyzer.

The Figure 7 illustrates the test system with a spectrum analyzer substituted for the power meter. A notch filter is also installed and is adjusted to suppress KFM an additional 25 dB, this insures that the analyzer front end is not overloaded. Note that the dynamic range is narrowed to approximately 40 dB; well within the 60 dB limit of an economical spectrum analyzer.

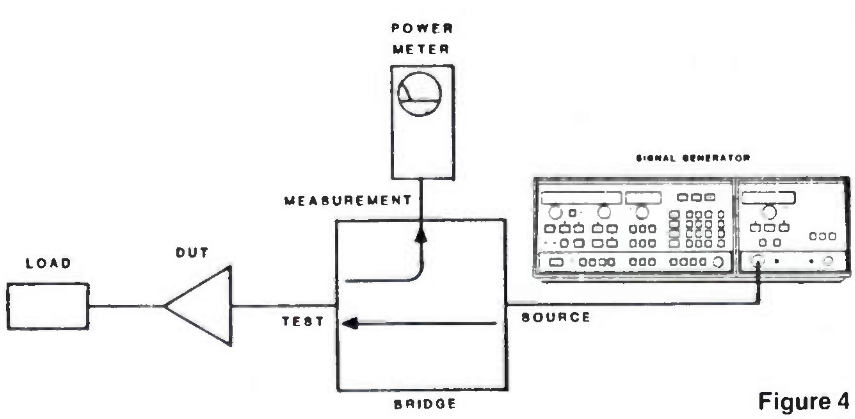


Figure 4

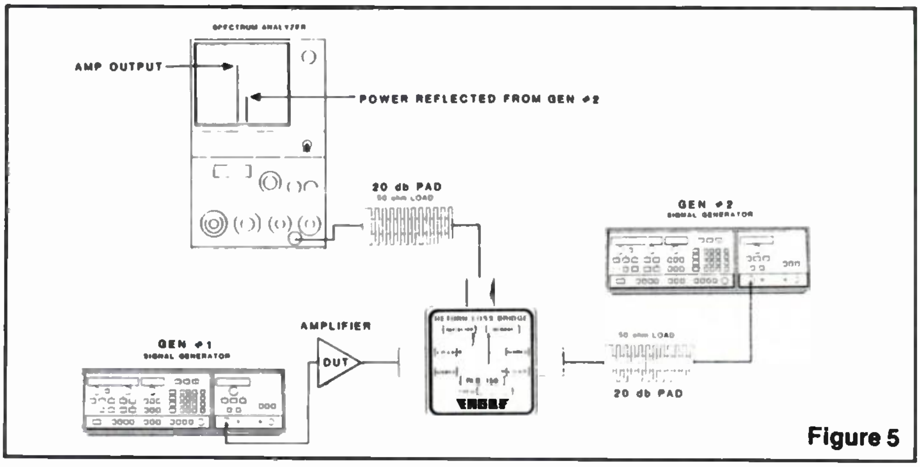


Figure 5

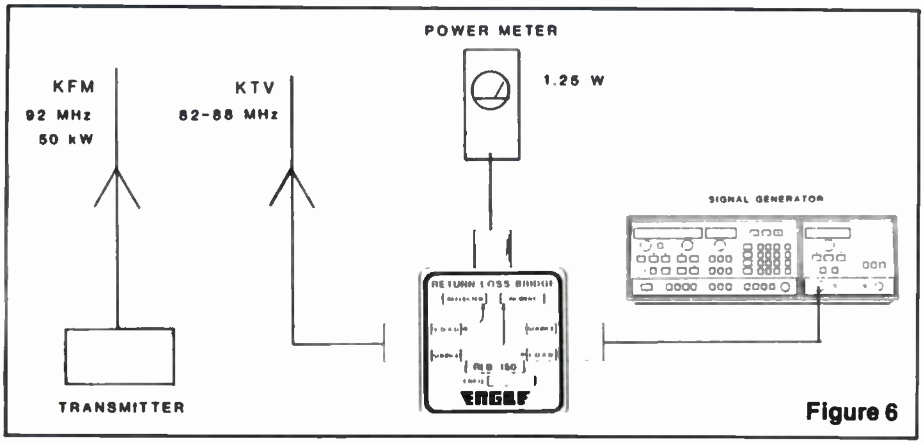


Figure 6

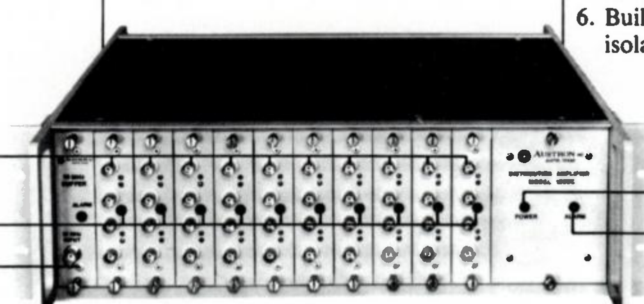
Mixer Port Return Loss and Isolation

The advent of high level diode mixers has brought with it associated problems in testing these devices. Parameter measurement is a time consuming and elaborate process that requires much care. The test system shown below will measure all important mixer parameters simultaneously, this includes port return loss, port to port isolation and intermodulation distortion. If sweep generators are used, this can be a swept test that covers the frequency range of the return loss bridge. Using this setup mixer can easily be tested in less than five minutes.

The test procedure is as follows:

1. Apply drive and RF power at desired levels. Read LO port return loss on SA 1. Since the return loss bridge incident (MEASUREMENT PORT) port is sourced by the mixer read RF feed through on SA 1. If desired LO power can be varied while return loss and feed through are monitored.

2. On SA 2 observe the RF port reflected power and feed through from the LO and IF ports of the DUT. To run intermodulation select generator



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SAS-200/512	200 - 1800 MHz	Log Periodic
SAS-200/518	1000 - 18000 MHz	Log Periodic
SAS-200/530	150 - 550 MHz	Broadband Dipole
SAS-200/540	20 - 300 MHz	Biconical
SAS-200/541	20 - 300 MHz	Bicon / Collapsible

MODEL #	FREQ. RESP.	DESCRIPTION
SAS-200/542	20 - 300 MHz	Biconical, Folding
SAS-200/550	.001 - 60 MHz	Active Monopole
SAS-200/560	per MIL-STD-461	Loop - Emission
SAS-200/561	per MIL-STD-461	Loop - Radiating
BCP-200/510	20 Hz - 1 MHz	LF Current Probe
BCP-200/511	100 KHz-100 MHz	HF/VHF Crnt. Probe

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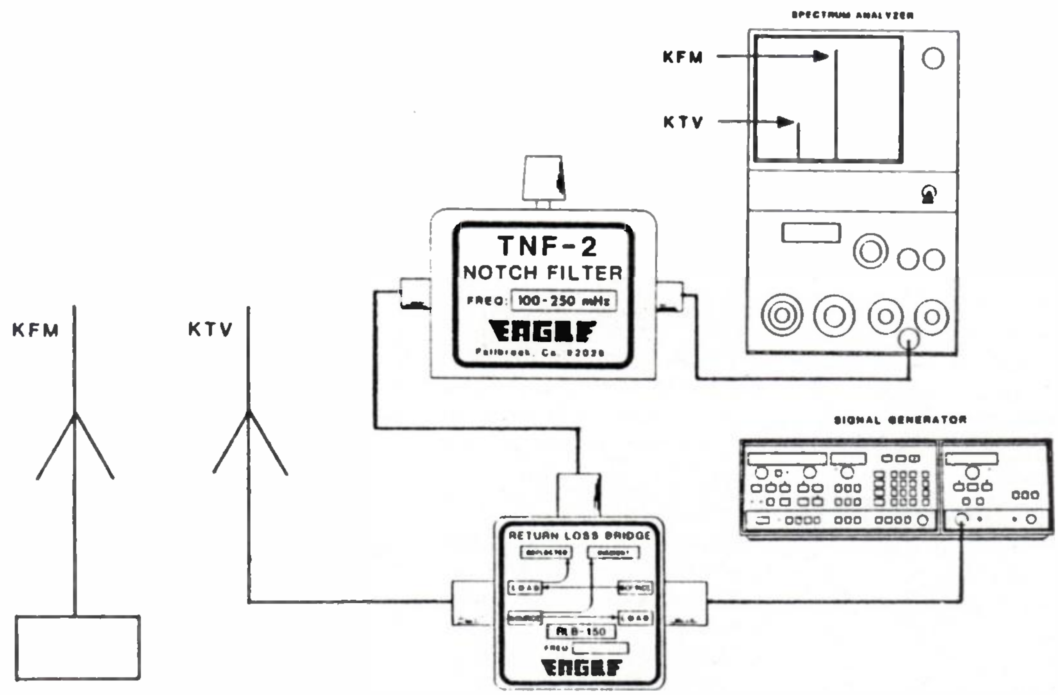


Figure 7

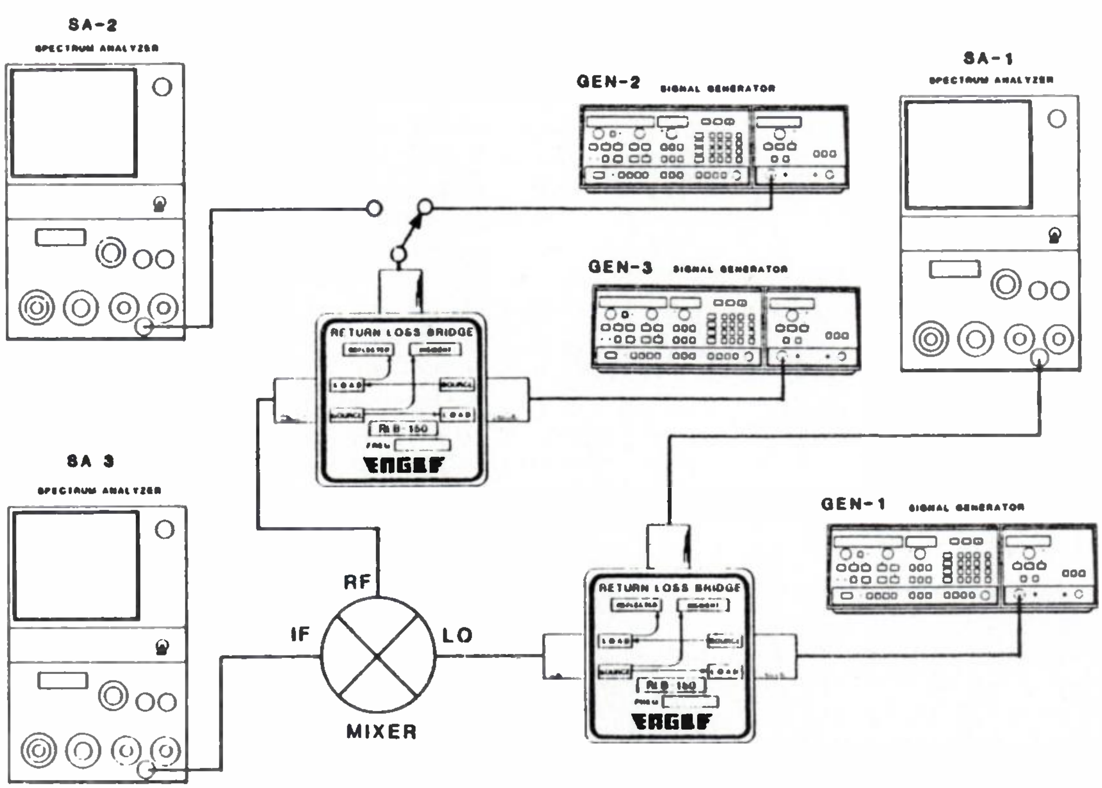


Figure 8

2 on RF port, the return loss bridge now becomes a power combiner thus supplying a two tone signal to the RF port of the DUT.
3. On spectrum analyzer 3 observe IF output, intermodulation distortion as well as feedthrough from the RF and LO ports of the mixer.

Naturally the bridges used for this system must have an RF MEASUREMENT PORT. A bridge with a five watt rating can easily supply the required drive levels to any diode type mixer.

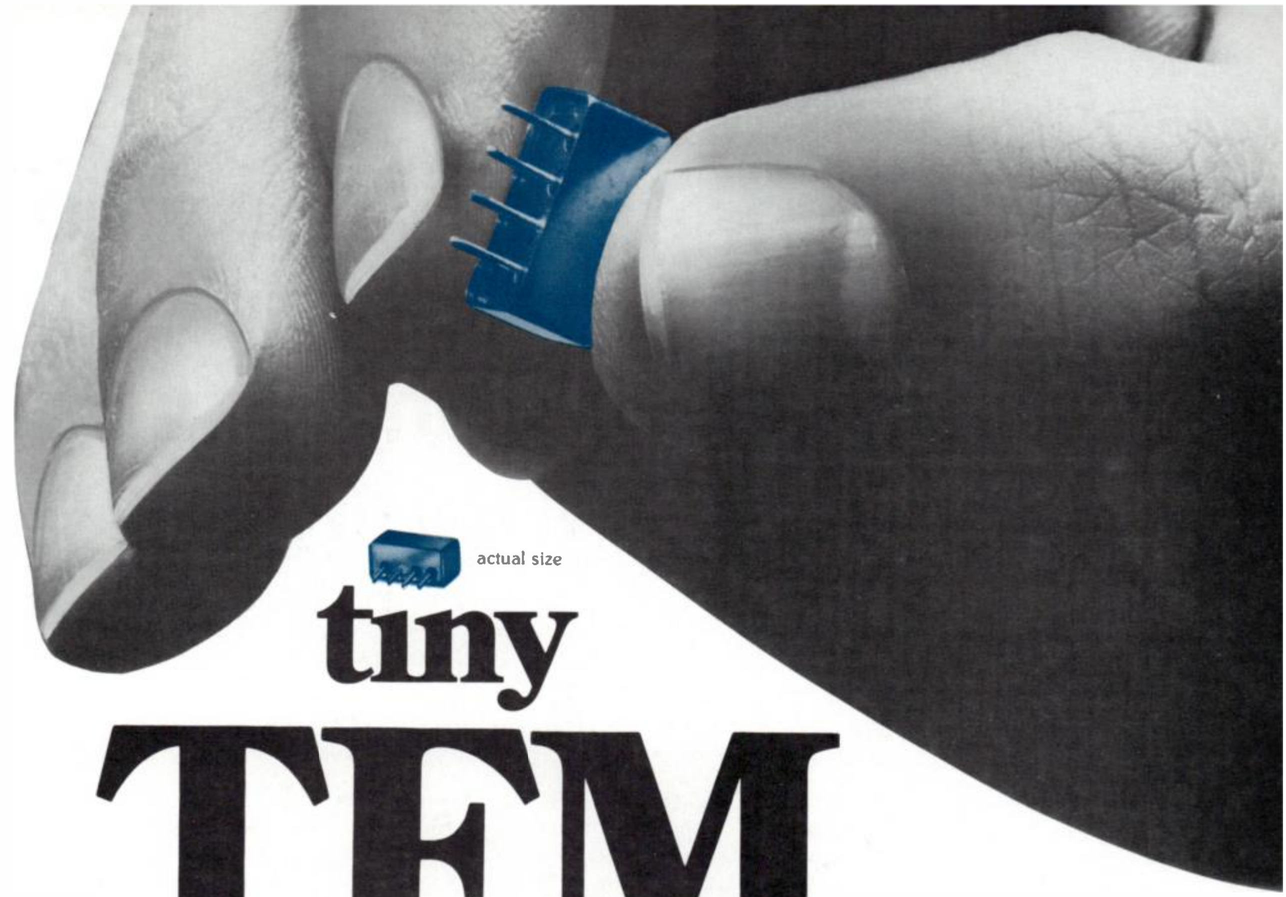
Impedance Using a Vector Network Analyzer

A vector network analyzer may be connected directly to the MEASUREMENT port of a return loss bridge provided that it has an RF type port. Both magnitude and phase of the reflection are present at this port. A reference signal can be derived by using an isolating power divider at the SOURCE PORT. The second output is then fed into the network analyzer as a reference signal. It is recommended that a resistive divider be used because of the excellent phase balance inherent in this device.

Summary

The Return Loss Bridge is a useful device for making low power precision VSWR measurements. Most of these devices will be able to yield a -40 dB directivity which is equivalent ot a 1.02 VSWR. Performance in excess of this is only possible using slotted line techniques. Return loss bridges can be used for characterizing passive networks, antennas and low power RF amplifiers.

Because of the size, price, high performance and ease of use the return loss bridge is, or should be, a standard piece of test equipment in the RF laboratory or service center.



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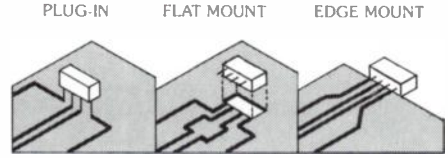

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MODEL	FREQUENCY, MHz		CONVERSION LOSS dB, TYPICAL		ISOLATION dB, TYPICAL						PRICE	
			1 Octave from Band Edge	Total Range	Lower Band Edge		Mid Range		Upper Band Edge			
	LO/RF	IF			LO-RF	LO-IF	LO-RF	LO-IF	LO-RF	LO-IF	$ EA.	QTY.
TFM-2	1-1000	DC-1000	6.0	7.0	50	45	40	35	30	25	11.95	(1-49)
TFM-3	.04-400	DC-400	5.3	6.0	60	55	50	45	35	35	19.95	(5-49)
TFM-4	5-1250	DC-1250	6.0	7.5	50	45	40	35	30	25	21.95	(5-49)
•••TFM-11	1-2000	5-600	7.0	7.5	50	45	35	27	25	25	39.95	(1-24)
•••TFM-12	800-1250	50-90	—	6.0	35	30	35	30	35	30	39. 95	(1-24)
••TFM-15	10-3000	10-800	6.3	6.5	35	30	35	30	35	30	49.95	(1-9)
••TFM-150	10-2000	DC-1000	6.0	6.5	32	33	35	30	35	30	39.95	(1-9)

•••If Port is not DC coupled
••+10 dBm LO, +5 dBm RF at 1dB compression
*Units are not QPL listed

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INFO/CARD 22

Reference-Level Control In Spectrum Analyzers: Simplification Through Internal Intelligence

By Morris Engelson
Tektronix, Inc.
Beaverton, Oregon

Both users and designers of spectrum analyzers are constantly seeking ways to simplify instrument operation. While a modern spectrum analyzer includes many control functions, three are essential: frequency tuning, frequency span across the screen, and amplitude level. Unfortunately, each of these functions involves several circuit operations, so that more than one control is usually required for each function.

Inclusion of increasingly more sophisticated circuits, coupled with microprocessor intelligence, has moved the spectrum analyzer closer and closer to a true three-knob-control instrument, For example, microprocessor monitoring of which local oscillator is to be tuned, depending on phaselock condition, has resulted in just a single center-frequency control, with internal switching among oscillators transparent to the user.

Changing the amplitude function into a single control is much more complicated. The solution chosen for the Tektronix 492, 494, and 496 Spectrum Analyzers departs from previous practice, although the control operation is quite similar to that used in the popular 7L5. The arrangement may seem strange to those not used to it, but once understood, it saves the user much extra control manipulation.

The Reference-Level Function

Amplitude-calibrated spectrum analyzers provide a logarithmic vertical display in dB/div, in addition to a linear volts/div scale. Because dB (decibels) is a relative amplitude measure, dB differences must be compared to a "reference level" to obtain an absolute measure of amplitude. Theoretically, the reference level could be anywhere on screen, although by convention it is the top graticule position. The reference level is defined as the absolute input signal level, in dBm, that produces full screen display.

The reference level is changed whenever instrument transmission gain is increased or decreased (attenuation). An increase in gain permits a smaller input signal to produce a full screen display; thus, the reference level is decreased (e.g., from -30 dBm to -40 dBm). A decrease in gain, or an increase in signal attenuation prior to the gain stages, means that the input signal level has to increase to produce a full screen display; thus, the reference level is increased (e.g., from -30 dBm to -20 dBm).

Clearly, there are two separate instrument areas that affect reference level — gain and signal attenuation. Ideally, it would be desirable to influence both of these areas with one control. Gain change or input signal attenuation can also reduce the signal level and create different instrument behavior for noise level and distortion. The user then needs to choose between these parameters. Furthermore, input impedance characteristics and mixer protection against large (especially pulsed) signals depend on the choice of input attenuation.

A Model

Circuitry for controlling the reference level in spectrum analyzers can be represented by the model shown in Figure 1. At some particular level of input attenuation (call it 0 dB), and

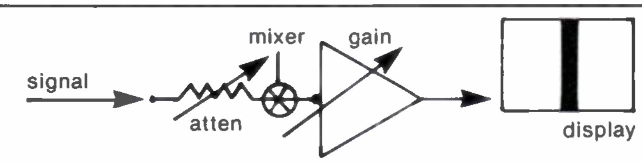
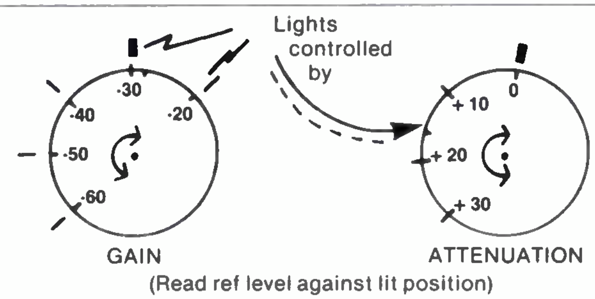
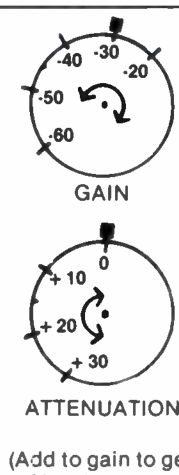


FIG. 1



(Read ref level against lit position)

FIG. 3



(Add to gain to get ref level)

FIG. 2

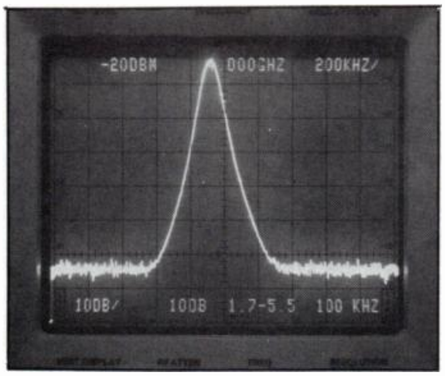


Figure 4

amplifier gain (call it 0 dB), the zero-reference input signal produces a full-screen reference-level display. The zero-reference signal is defined as the input level which, on the whole, provides the best dynamic measurement range as noise, distortion products, and on-screen display capability are balanced against each other. High amplifier gain increases internal noise and thus reduces the measurement range, while very low gain requires high input signal levels into the mixer and results in measurement-limiting distortion products. There is no one absolutely best zero level under all measurement conditions. The range runs from about -20 dBm to -40 dBm, with most spectrum analyzers designed around a -30 dBm level. A zero level of -30 dBm is used in the 490 Series.

How It Was Done Before

In the earlier years of calibrated spectrum analyzers, reference levels were set by two independent controls: one for gain (or IF gain), and another for input (or RF) attenuation., Both controls were calibrated in dB, with zero corresponding to the zero-reference input signal level. Knowing the zero level (e.g., -30 dBm), the user could add or subtract gain or attenuation to compute the reference level. In a more sophisticated variation of this method, the zero-reference gain position was labeled at -30 dBm, as shown in Figure 2. Gain changes would then indicate the reference level directly, so long as attenuation was not changed.

A later improvement provided a moving reference indicator around the gain control, as shown in Figure 3. The indicator was either mechanically or electrically (lights) coupled to the attenuator control. This method eliminated the need for the mental arithmetic of the previous approach.

Another version provided for a mechanical interlock between the two controls so that an indicator window or other mechanism would move to provide for direct readout of reference level.

Whatever the arrangement, the user had to manipulate two controls to get the desired reference level because gain and attenuation were independent variables, with reference level dependent. In the 490 Series, however, reference level is set directly, while gain and attenuation change as needed to accommodate the desired reference level. This solution results in a single control arrangement.

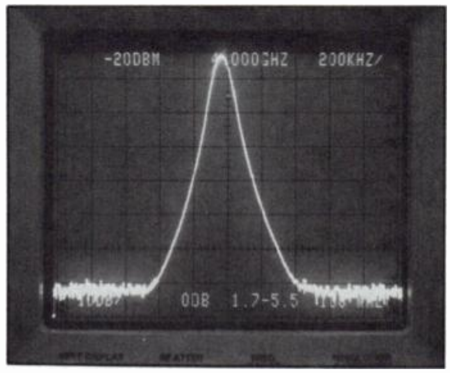


Figure 5

The Tektronix Arrangement

Reference level (R in dBm), gain (G in dB), and attenuation (A in dB) are related through the following simple equation: $R = G + A + C$, where C is a normalization constant. Clearly, the same value of R can be obtained through more than one combination of G and A. Therefore, it is necessary to have independent access to either gain or attenuation even when the reference level can be manipulated directly. In the 490 Series, such access is provided by providing a "minimum attenuation" control that sets the minimum level of attentuation below which the instrument will not go. (This control should not be confused with the attentuation setting control in previous arrangements.) Once the "min-atten" control is set, the reference

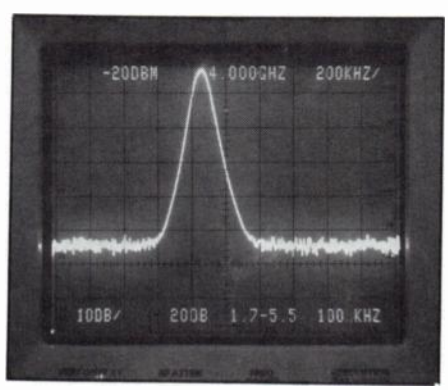


Figure 6

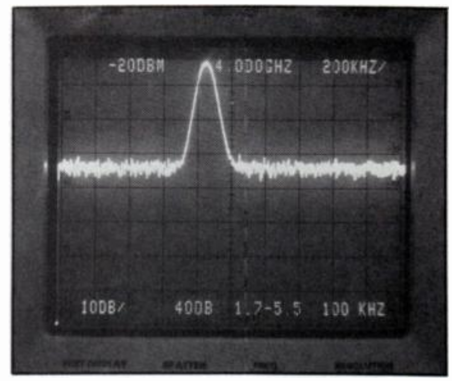


Figure 7

level is actuated by a single control that switches in gain or attenuation as needed to provide the desired reference level. The choice of gain and attenuation is determined by two restrictions: attenuation cannot go below the minimum chosen, and gain will not be reduced below the point that provides a good dynamic range.

Trading Dynamic Range For Noise Reduction

A signal with an amplitude greater than the reference level will produce an off-screen display. The spectrum can be brought on-screen by reducing the apparent signal level in either of two ways, front-end attenuation or gain reduction. Gain reduction reduces the displayed noise level and thus permits more of the screen to be used for signal display. However, the front-end spectrum analyzer circuits are then subjected to greater signal levels and produce more distortion (compared to the attenuation technique of signal-level reduction). In the 490 Series, the choice of operating conditions is provided by a min-distortion/min-noise control The minimum noise setting provides 10 dB of noise reduction instead of 10 dB front-end attenuation (although attenuation will still not go below the minimum setting).

Although other spectrum analyzers provide 20 dB, 30 dB, or even as much as 40 dB of gain reduction, a limit of 10 dB is used in the 490 Series to assure that the on-screen display reflects real signals rather than spurious responses. This choice (of how much gain reduction to provide) is not related to how the gain attenuation and reference level are controlled, but rather to the fact that the intermodulation dynamic range drops by at least twice the rate of gain reduction. For example, a 40-dB gain reduction cuts an 80-dB dynamic range to zero — not a particularly useful measurement arrangement. Reductions of 10 dB or 20 dB do provide a reasonable intermod-

The 490 Series Arrangement Compared To Others

Advantages

- Permits presetting the minimum attenuation desired. Set it, then forget it.
- True single knob reference level control. No need to manipulate two controls to get the desired ref. level.
- Choice of minimum distortion or minimum gain algorithms. The user knows what he is getting. Comes up the same way (minimum distortion) every time the instrument is turned on.
- Does not provide too much gain reduction to avoid mistaking spurious responses for real signals. Additional gain reduction can be simulated by going off screen.

Disadvantages

- New scheme, people are not used to it.
- Noise level will be high and stay high if the min-atten is set at a high level (e.g. 40 dB), and the user does not realize it.
- Not obvious how to get a desired gain and attenuation setting as opposed to a reference level.
- Only 10 dB of gain reduction means that for some rare applications one needs to drive the signal off screen.

ulation measurement range. Each has its merits and disadvantages, and the choice is pretty subjective. In any event, it is always possible to drive the signal off screen, thus simulating a larger gain reduction.

Advantages and disadvantages of the arrangement used in the 490 series are listed in the table above.

Photographic Examples

The vertical calibration arrangement provided by the 490 series permits setting all combinations of gain, attenuation, and reference level. Figure 4 shows the normal setting intended for best dynamic range, which occurs at -30 dBm for 0 dB front-end attenuation. The microprocessor sets 10 dB of attenuation to reach -20 dBm reference, although the minimum attenuation was preset at 0 dB.

Figure 5 shows what happens when the minimum noise position is chosen. The 10 dB of rf attenuation is removed, and gain (also on-screen noise) drops by 10 dB. The reference level is unchanged, and the full-screen signal display remains full-screen.

Figure 6 shows what happens when the minimum attenuation is preset to 20 dB. Gain and noise increase, attenuation is up to 20 dB, and the reference level and signal display amplitude are unchanged.

Figure 7 shows a similar situation at 40 dB minimum attenuation. As stated earlier, all possible combinations of gain (normal, -10 dB, and increase by over 100 dB in 1-dB steps), attenuation (zero to 60 dB in 10-dB steps), and reference level can be set with the control system in the 492, 494, and 496.

Although the 490 Series provide a maximum 10 dB of gain (noise) reduction, the equivalent of greater reduction can be achieved by driving the signal off-screen. This approach has the advantage that most on-screen displays are real; thus, there is no possibility of erroneously identifying a spurious response as a real signal. The disadvantage is that on-screen displays cannot be achieved on those occasions when noise reduction greater than 10 dB is desired.

Figure 8 shows a normal two-signal display. Intermodulation products are below noise level.

Figure 9 shows the same two 0-dBm

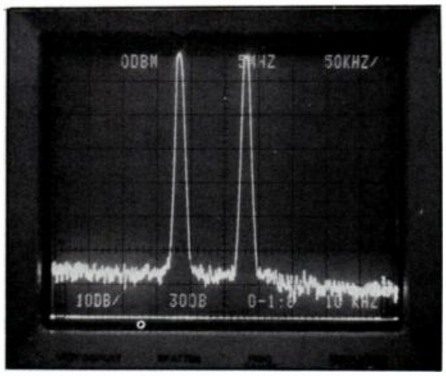


Figure 8

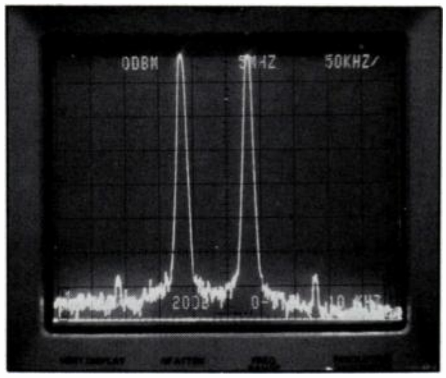


Figure 9

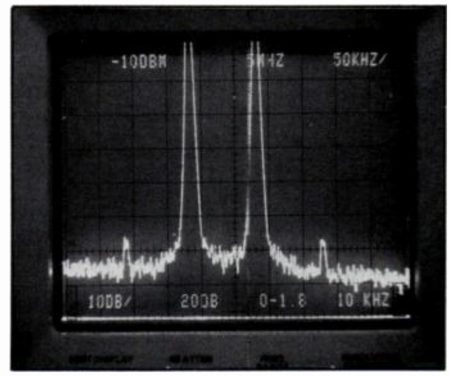


Figure 10

signals in the minimum noise mode. The noise level has dropped by 10 dB, and the input circuits are subjected to 10 dB greater signal level as the front-end attenuator is down from 30 dB to 20 dB. Third-order intermodulation products can now be observed on the screen. The information provided by Figure 9 can also be obtained by changing the reference level by 10 dB and driving the signal 10 dB off-screen, as shown in Figure 10.

Going another 10 dB off-screen brings the third-order products up to only 50 dB below the input signals, as shown in Figure 11, a composite of two photos. Finally, Figure 12 shows the display on a competitive instrument, where gain reduction greater than 10 dB is possible. Although the signal spectrum is all on-screen, the intermodulation products are not real. Figure 8 shows that these products are significantly more than 50 dB down.

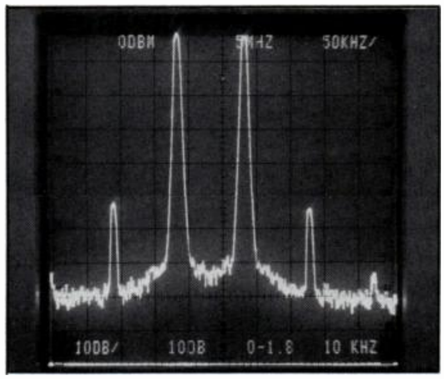


Figure 11

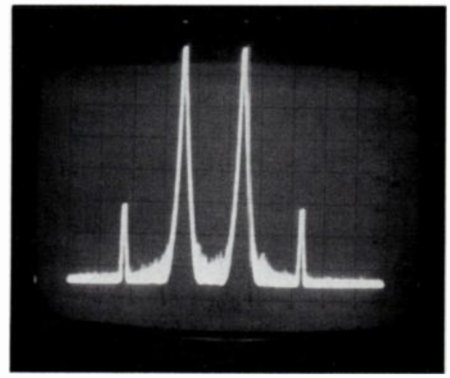


Figure 12

Conclusion

Perhaps one day, artificial intelligence will reach a level where human judgment will be replaced by machines. We are certainly not there now. Therefore, we continue to have some degree of ambiguity in applications involving complicated interactions or relationships. The vertical display level of the spectrum analyzer is no exception.

Current efforts are in the direction of assuring that the more common measurements can be done more or less automatically; that the control arrangement is as simple to use and learn as possible and that the human operator knows when to, and is able to, override the computer.

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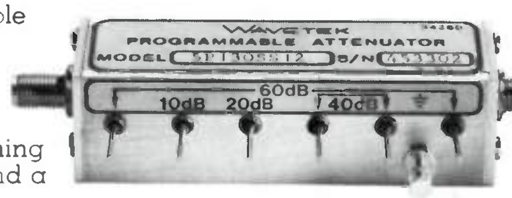

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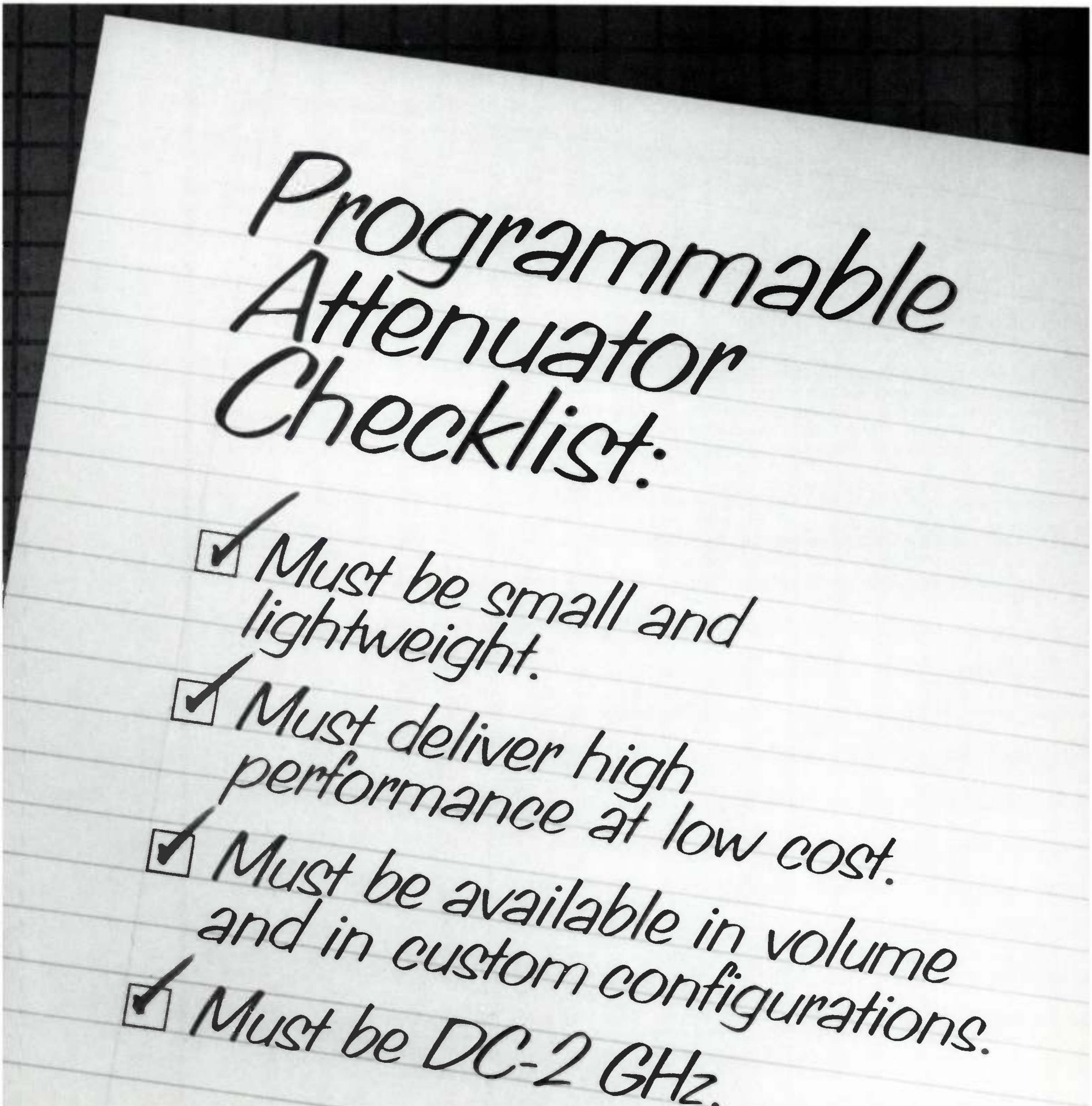

DEMONSTRATION INFO/CARD 36
LITERATURE INFO/CARD 86

"Near Field" Communication

The "Near Field" low frequency communication applications have been neglected for quite some time. However, suitable applications still exist and should be more closely examined.

By Andrzej B. Przedpelski
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A.R.F. Products, Inc.
Boulder, CO 80301

Despite the fact that "microwaves" became quite a buzz-word in the last few decades, there are still some applications for the lowly low frequencies. In the thirties, high power low frequency broadcasting stations were quite de rigueur practically everywhere but in the United States. I remember, as a child, putting together my first crystal set to listen to Warsaw I at about 200 kHz. Thanks to the high power output and a good ground wave, reliable reception could be had with the simplest equipment. In those days all you needed was a crystal (not a JANTX diode, but a hunk of galena with a "sensitive spot"), some wire, a cold pipe for ground and the power line for an antenna (if you were very careful - I almost burned the house down when my "antenna" shorted to my "ground") and you were in business. During World War II, when amateur transmissions were restricted, low frequencies enjoyed a short revival in the United States. Some hams, not willing to give up their hobby, switched to carrier current as means of communicating. This meant operation below 100 kHz, using the power lines as transmission lines. Theoretically, the radiation was supposed to be negligible, but it was possible to "jump" from one power line to another. This was handy in places like Boston, where some of the power was AC and some DC therefore line jumping was necessary to obtain reasonable distances.

Long range (high power) applications for low frequencies always existed, such as: communicating with submarines at sea, navigation (Omega, Loran-C, Decca), and timing (WWVB and US Navy standard frequency and time stations such as NAA, NSS NLK).

For a while things were quiet on the low frequency front, until garage door openers first became popular in the late forties. To provide "license-less" operation (it is doubtful if the average motorist could pass any kind of FCC exam), a very low power operation was required by the FCC. This is where "near field" operation comes in [Reference 1]. To meet the low maximum allowable field strength permitted, the fact that the field varies as the inverse cube of the antenna separation for distances considerably less than a wavelength, was utilized. At distances over one wavelength, the usual inverse law applies. Thus, using very low frequencies (long wavelengths), a usable signal level could be obtained while still satisfying the FCC requirements.

It may be desirable, at this point, to define what is meant by "near field." An RF current in a transmitter loop antenna produces an oscillatory magnetic field. This field, in turn, induces a proportional current in a receiver loop antenna placed in it. Thus, the transmitter loop — receiver loop system can be thought of as a very loosely coupled ("leaky") transformer (see Figure1).

The formula for mutual inductance of two parallel coils, as shown in Figure 1, is [Reference 2]:

$$M = \frac{\mu.\pi}{16} \sqrt{r_R r_T} \left[\frac{4\, r_R r_T}{(r_R + r_T)^2 + d^2} \right] \tag{1}$$

which, for $r_R + r_T << d$ (the usual case), can be simplified to:

$$M = 0.2\pi^2 \frac{r_T^2 r_R^2}{d^3} \tag{2}$$

Using the formula for induced voltage:

$$e = -j\omega Mi \tag{3}$$

and taking into account the fact that multiturn coils are used, the overall system formula for induced voltage in the receiver loop antenna becomes:

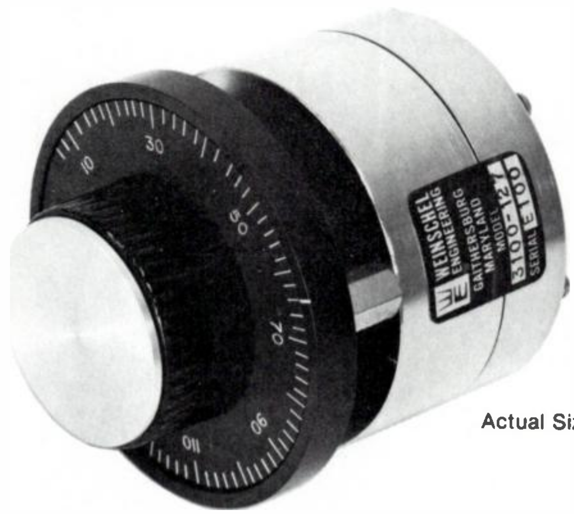
$$e_R = 0.4\pi^3 \frac{r_T^2 r_R^2 n_T n_R}{d^3} f\, i_T \tag{4}$$

Basically, the same results can be obtained using Henney's formula for radial near field [Reference 3]:

$$e_n = \frac{18.85\, n_T r_T^2}{d^3}\, i_T \times 10^7 \tag{5}$$

IFR, Inc. Model FM/AM 1500 Communications Services Monitor uses Weinschel Model 3100-127 to achieve a high degree of accuracy and dependability

Weinschel's 3100 Variable Attenuator Chosen by IFR for their FM/AM 1500



Actual Size

IFR, the leading designer and manufacturer of precision test equipment for the Avionics Electronics and Communications industry, chose Weinschel's 3100-127 variable attenuator for use in their FM/AM 1500 Communications Services Monitor. A member of the National Conference of Standards Laboratories, IFR demands a high quality and dependable OEM product.

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and Terman's for induced receiver loop voltage [Reference 4]:

$$e_R = 2\pi e N \frac{A}{\lambda} \quad \text{where } e \text{ can be } e_n \text{ or } e_f. \tag{6}$$

The induced receiver antenna voltage thus becomes:

$$e_R = 3.9585\pi \frac{r_T^2 r_R^2 n_T n_R}{d^3} f\, i_T \tag{7}$$

For comparison, the "far" field transfer function was derived:

$$e_R = 8.89\pi^5 \frac{r_T^2 r_R^2 n_T n_R}{d} f^3 i_T \times 10^{-18} \tag{8}$$

using [Reference 5]:

$$e_f = \frac{120\pi^3 n_T r_T^2}{d\lambda^2} i_T \tag{9}$$

and Equation (6).

The inverse and inverse cube relationships are clearly visible in Equations (8) and (7). Figure 2 shows them for normalized conditions ($r_R = r_T = n_R = n_T = i_T = 1$) for an RF frequency of 100 kHz. It can be seen that at a distance of just under ¼ wavelength the performance of the two approaches is equal. At shorter distances, the "near field" is clearly superior. At longer distances, the more common "far field" performance is better.

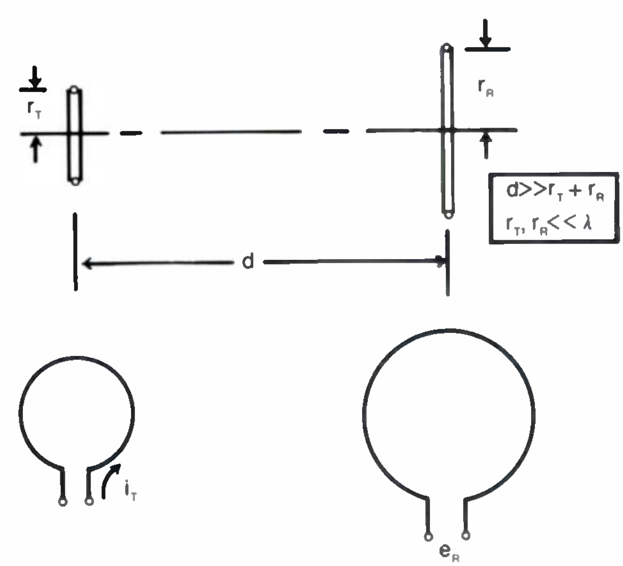

FIG. 1 Two Coaxial Parallel Loop Antennas


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3rd ANNUAL
Test&Measurement World EXPO
APRIL 3-5, 1984
BROOKS HALL, SAN FRANCISCO

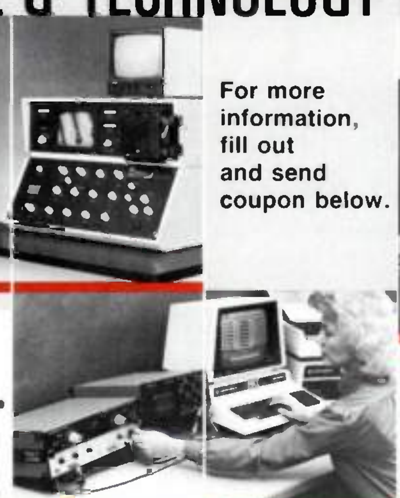

Table 1
Definitions

f = frequency, hertz
d = loop antenna separation, meters
n_T = number of turns in transmitter loop antenna
n_R = number of turns in receiver loop antenna
r_T = radius of transmitter and antenna loop, meters
r_R = radius of receiver antenna loop, meters
i_T = transmitter RF loop current, amperes
e_n = near-field strength, microvolts per meter
e_f = far-field strength, microvolts per meter
e_R = voltage induced in receiver loop antenna, microvolts
M = mutual inductance, microhenries
μ_o = permeability of free space, $4\pi \times 10^{-7}$ henries per meter
λ = wavelength, meters
A = loop antenna area, square meters

Table 2
FCC Regulations - Part 15

Maximum allowable field strength for nonlicensed operation:

Frequency - kHz	Distance - meters	Field Strength
10-490	300	2400/frequency
510-1600	30	24,000/frequency
Frequency - MHz		
70-130	30	50
130-174	30	50-150 (linear interpolation)
174	30	150
260-470	30	150-500 (linear interpolation)
470 and above	30	500

There are many applications where the "near field" approach provides important advantages. They are basically low power short range systems which can be divided into two main categories:

-systems which cannot exceed a certain field strength at a specified distance, and

-systems where partial security of transmission is desired.

A typical example of the first is the FCC low power requirement mentioned before in connection with the garage openers. To operate without a license, the low power FCC requirements have to be satisfied, as shown in Table 2. Figure 3 shows a comparison of operation at two frequencies: 100 kHz and 200 MHz. The maximum FCC allowable signal strengths are indicated. At distances of over about 400 meters the 200 MHz system will provide higher field strength, but at very short distances over 40 dB higher field strengths are provided using the low frequency approach. A free bonus is the reduced interference with other receivers located nearby, since the inverse cube relation applies to a distance of about 670 meters.

In addition to the above advantages, the lower frequency usually provides better propagation characteristics. It penetrates solid objects better and does not exhibit the usual nulls encountered in VHF operation.

The other application is in partially secure (or non-interfering) short distance communication systems. Let's assume

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that a system is needed having a minimum signal-to-noise ratio of 40 dB at transmitter/receiver separation of 100 meters. The intelligence bandwidth is 10 kHz. Two systems will be compared: one at 400 MHz and one at 50 kHz. Assuming a receiver noise level equivalent to a field strength of one microvolt per meter [Reference 6], a field strength of 100 microvolts per meter at 100 meters is needed, as shown in Figure 4. At a distance of 10,000 meters the system will have a 0 dB S/N.

On the other hand, the 50 kHz system, while requiring considerably higher field strength to provide the required S/N, will be lost in the atmospheric noise [Reference 6] (S/N of 0 dB) at a distance of only about 460 meters, as shown in Figure 5. Thus, security is considerably enhanced compared to the VHF system.

The necessity of using loop antennas is sometimes considered a disadvantage of the low frequency, near field systems. While modern loop design data is somewhat scarce, it is possible to design efficient structures [Reference 7] with minimal effort.

The above analysis is not meant to be all-inclusive, but it shows that the long neglected low frequencies still have some modern applications. Each individual short-range system requirements have to be considered before a final decision can be made, but in many cases the low frequency approach may have considerable advantages. In addition, after working at VHF and UHF it is quite a revelation to find out that capacitors really behave like the book says and an inductor is + j and not -j.

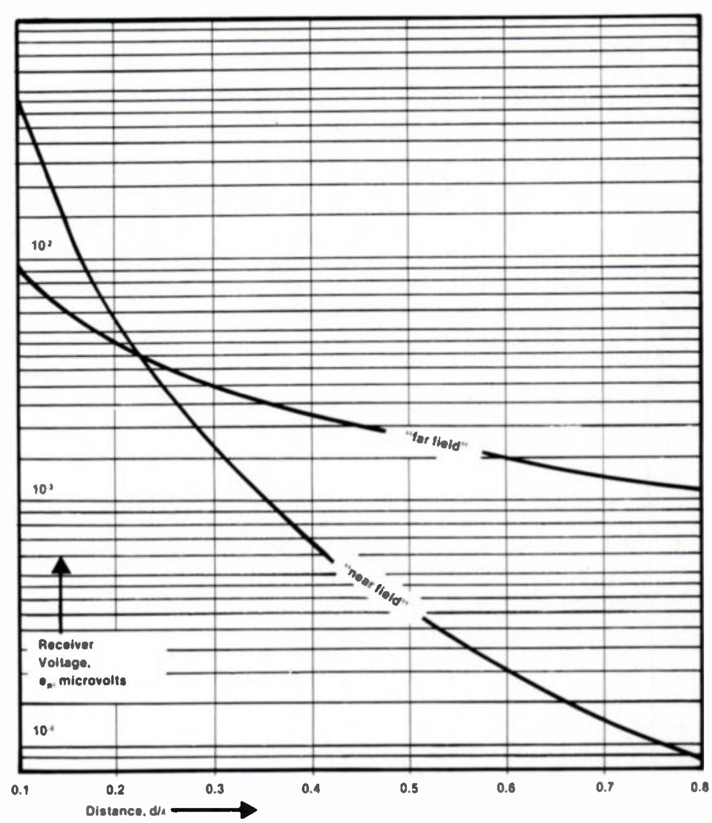

FIG. 2 Induced Receiver Voltage, e_R (Normalized)

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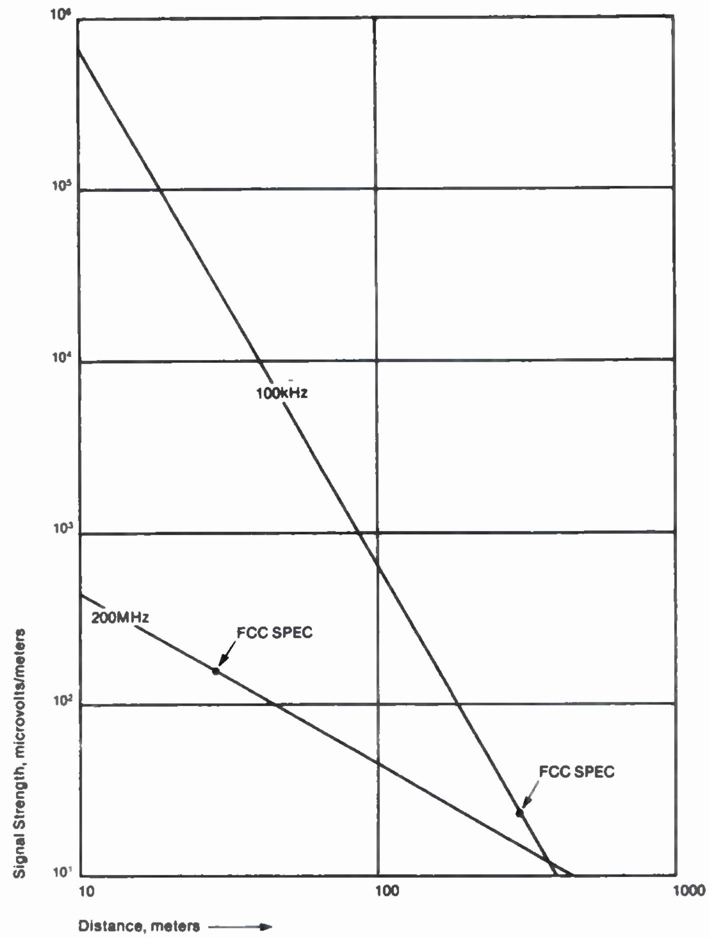

Fig. 3 Operation Under FCC "Low Power" Rules.

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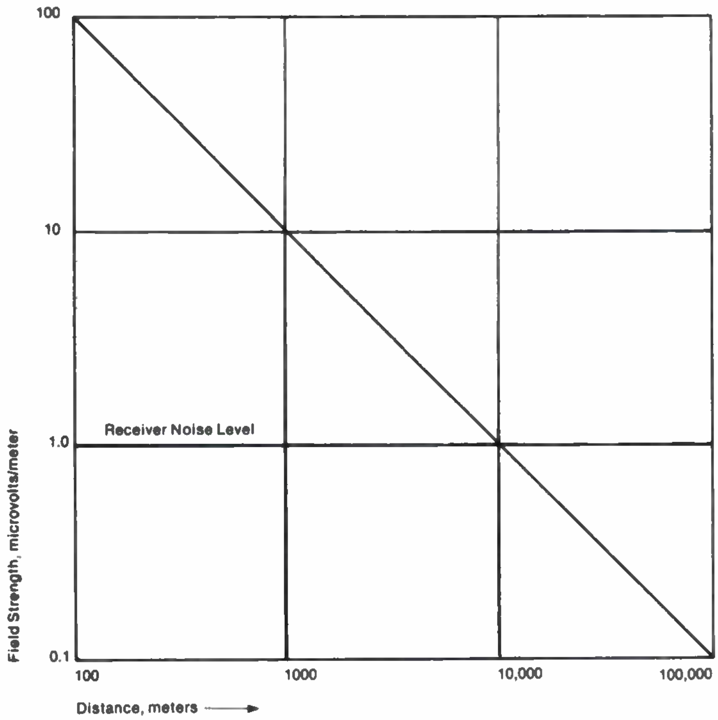

Fig. 4 Signal Levels Using 400 MHz.

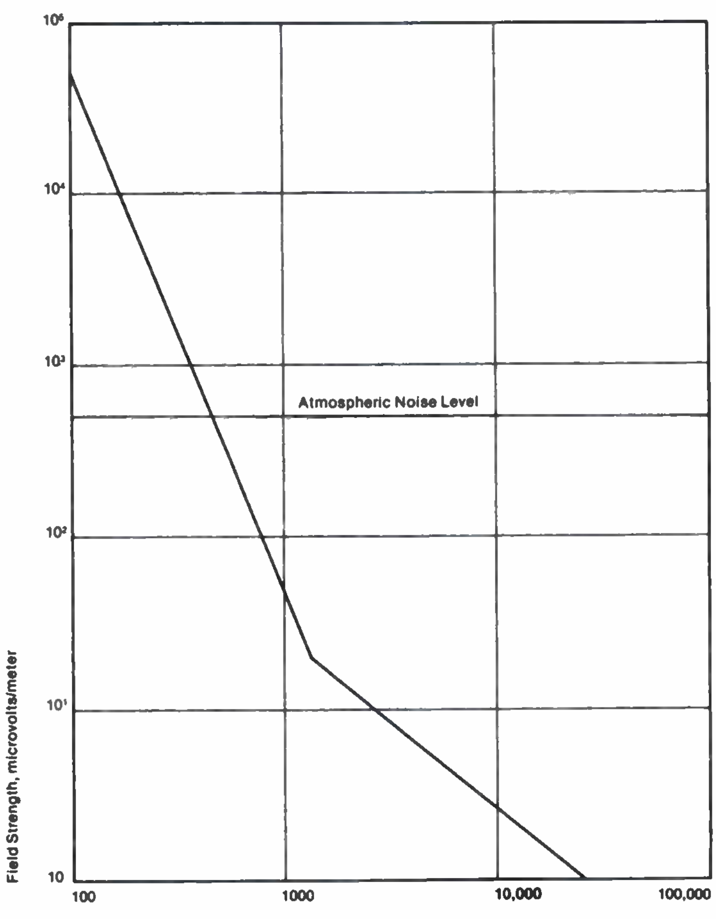

Fig. 5 Signal Levels Using 50 kHz.

Divider Delay: The Missing PLL Analysis Ingredient

This article will analyze PLL divider delay using control systems models and demonstrates the impact of delay on loop stability.

By Stan Goldman
Texas Instruments
Dallas, Texas 75230

Phase Lock Loops are used in many applications, but frequency generation is the most common. As design requirements for Phase Lock Loops (PLL) become more stringent, so does the importance of accurate analysis for loop performance. Most of the demanding PLL design decisions are related to increasing loop bandwidth to reduce noise and improve loop switching speed. The component values calculated using classical loop analysis equations have been used in circuits where the measured electrical response was not always in agreement with the calculated electrical response.

The discrepency between theory and experiment was found to be attributed to divider delay which caused a decrease in phase margin significant enough in many cases to cause unstable loop performance. Divider Delay is one of the most significant parameters which limits larger Phase Lock Loop bandwidths. It is defined as the time delay from a changing frequency input to the dividers until the detection of the changing frequency by the phase detector. The time delay results from the many clock pulses required to change the divider's output state which is sampled by the phase detector's input reference clock frequency rising or falling edge. This article will analyze Divider Delay using control systems models and demonstrate the impact of delay on loop stability.

To begin the analysis, Control Systems Theory will be reviewed, followed by the derivation of a PLL control system model. Next, through several algebraic rearrangements, equations will be derived to solve for circuit elements from which Bode plots of magnitude and phase of the Open Loop Transfer Function can be made. Finally, an example will be used to show the effects of Divider Delay.

The generalized Phase Lock Loop Function can be expressed in terms of Control Systems Theory as follows:

$$\frac{\theta_o(s)}{\theta_i(s)} = \frac{G(s)}{1 + G(s)\,H(s)} \qquad (1)$$

Stability is determined by the G(s) H(s) function, which is referred to as the Open Loop Gain Function. If G(s) H(s) ever becomes equal to -1, then equation 1 becomes unstable. Naturally, the parameters that define the G(s) H(s) function must be carefully scrutinized.

A type 2 second order loop is used to illustrate the analysis technique as shown in Figure 1. This circuit con-

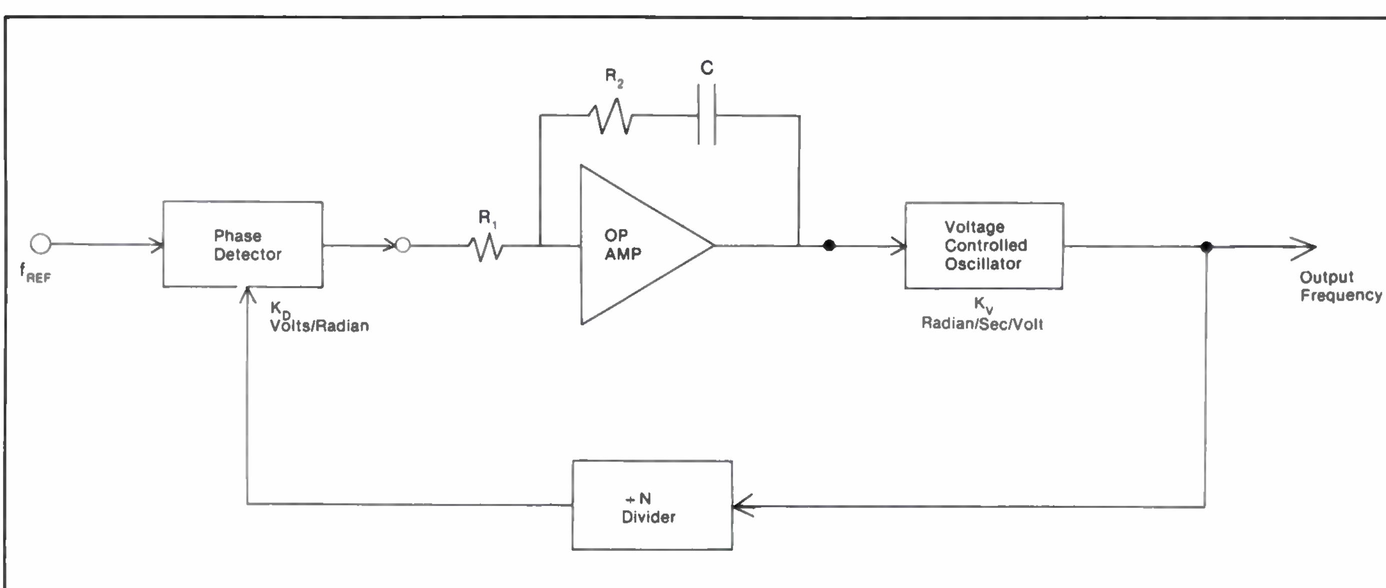

Figure 1. A Type 2 Second Order Loop.

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figuration is generally used in frequency generation applications. From Figure 1 the Open Loop Gain Function can be mathematically described as follows:

GH(s) = (Phase Detector Gain) (Filter Transfer Function) X (VCO Transfer Function) (Divider Transfer Function) (2)

$$= (K_D)\left(\frac{sCR_2+1}{sCR_1}\right)\left(\frac{K_V}{s}\right)\left(\frac{1}{N}\right)\left(e^{-sT}\right) \qquad (3)$$

$$= \left(\frac{K_DK_V}{NCR_1}\right)\left(\frac{1}{s^2}\right)\left(sCR_2+1\right)\left(e^{-sT}\right) \qquad (4)$$

where:

T = 1/Reference Frequency $\quad K_V$ = VCO Transfer Function
K_D = Phase Detector Gain $\quad$ N = Integer divider value

Equation 4, the Open Loop Transfer Function, is organized into terms which can be expressed easily using Bode Plot graphic techniques. The first term represents a gain constant; the other terms represent Bode Plot functions of frequency. Notice the two poles at the origin. This term represents two integrators which will force the error function closer to zero with decreasing frequency. The last term is the mathematical representation of delay. Equation 4 was derived using references 2 and 3.

When designing a Phase Lock Loop, the loop requirements are generally specified by loop bandwidth, output frequency (N x Reference Frequency), and stability (Damping Factor) parameters. Using these parameters and equation 4 one can determine circuit elements necessary to meet the loop performance requirements.

To calculate circuit elements for the PLL requires several operations. First, an equation for phase variations versus frequency must be developed and used to solve for the frequency location of the Zero ($1/2\pi R_2C$). Next, a magnitude versus frequency equation must be calculated and used to solve for the value of the Gain Constant term (K_DK_V / NCR_1).

Stability requirements are determined by the phase of GH(s). If the phase equals 180° and the magnitude equals unity, then oscillations will occur. The loop bandwidth is the frequency where the Open Loop Transfer Function equals unity. The last term of equation 4, ($(s\,CR_2+1)e^{-sT}$) contains phase variations versus frequency, while the other terms have a constant phase. Therefore, the other terms can be put aside. Now, solving the last term in equation 4 for phase:

$$\text{Phase} = \text{ARCTAN}\left[\frac{(\sin(-\omega T) + (\omega R_2C)(\cos(-\omega T))}{\cos(-\omega T) - (\omega R_2C)(\sin(-\omega T))}\right] \qquad (5)$$

Where:

$\omega = 2\pi f$

f = Frequency variable

Equation 5 can be used to plot the Open Loop Transfer Function's phase variation as a function of frequency. To find the frequency location of the Zero and thereby insure the required loop stability, equation 5 has to be rearranged, and values substituted. The stability requirement can be expressed by Damping Factor values and then expressed in terms of phase by the following formula:

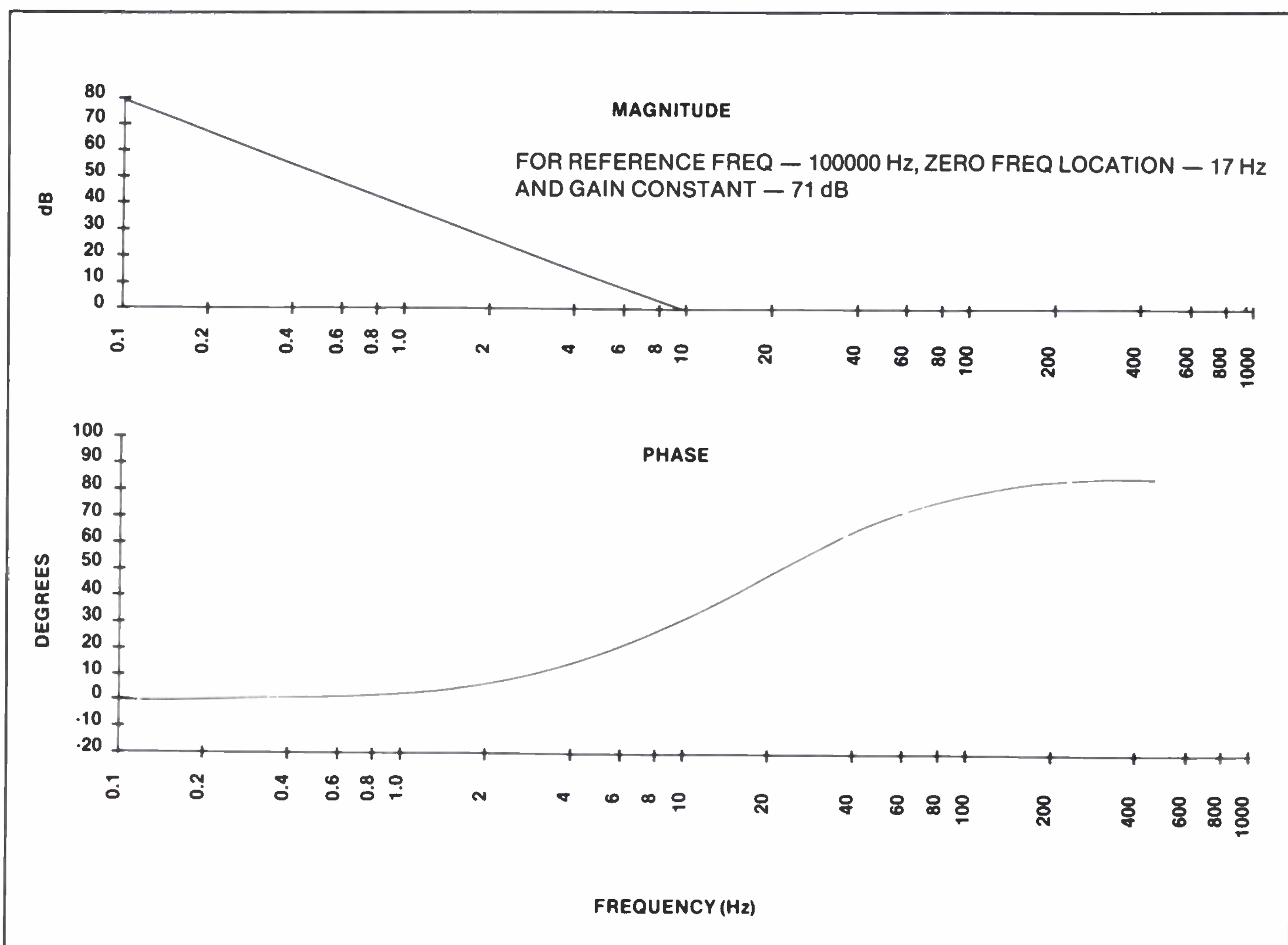

Figure 2. Classic PLL Solution

Phase Angle = 2DX

where:

D = Damping Factor

$$X = (2D^2 + (4D^4 + 1)^{0.5})^{0.5} = \frac{\omega_{BW}}{\omega_N}$$

ω_N = natural frequency

$$\frac{f_{BW}}{f_{BW(3dB)}} = \frac{X}{(1 + 2D^2 + ((1 + 2D^2)^2 + 1)^{0.5})^{0.5}} \approx 1$$

EXAMPLE
Optimum Solution of Figure 4

Given:	Detector	= .4V/Radian (Fairchild 11C44)
	K_V	= 125,000 (XR 2207)
	N	= 64
	D	= .27
	BW	= 10 Hz
	Ref. Freq.	= 100 Hz
Choose:	C	= 1 ufd
Solution:	R_1	= 489K
	R_2	= 35.9K
	BW (3dB)	= 15 Hz
	Freq. Crossing	= 10 Hz
	R_2/R_1	= .0735

Now, the phase angle value where the magnitude of the Open Loop Transfer Function equals unity determines phase stability. The number of phase angle degrees greater than -180 degrees (ie. instability) is called the Phase Margin. The frequency where the Open Loop Transfer Function equals unity is defined as the Loop Bandwidth. Therefore, equation 5 can be solved for the Zero Frequency Location:

$$f_{zero} = \frac{\left[(-\omega_{BW}\sin(\omega_{BW}T))(\mathrm{TAN}(\text{phase angle}) + \omega_{BW}\mathrm{Cos}(\omega_{BW}T)\right.}{\left[(\mathrm{TAN}(\text{phase angle}))(\cos(\omega_{BW}T) + s(\omega_{BW}T)\right]2\pi}$$

where:

$$f_{zero} = \frac{1}{2\pi R_2 C}$$

$$\omega_{BW} = 2\pi f_{BW(3dB)}\left(\frac{f_{BW}}{f_{BW(3dB)}}\right) = \text{zero magnitude crossing of GH}$$

T = 1/Reference Frequency = N/Output Frequency
N = Loop Frequency Multiplication Factor

f_{BW} = Loop Bandwidth (Zero Crossing)

$f_{BW(3dB)}$ = 3dB Bandwidth

Now that the frequency location of the Zero is determined, more circuit parameters can be calculated. The magnitude response of equation 4 can now be solved:

$$|GH(f)| = \left(\frac{K_D K_V}{NCR_1}\right)\left(\frac{1}{\omega^2}\right)\left[\left(1 + \left(\frac{f}{f_{zero}}\right)^2\right]^{0.5} \quad (8)$$

where:

$\omega = 2\pi f$

Equation (8) can be used to plot the Open-Loop Transfer Function magnitude versus frequency. However, since the Zero frequency is located by solving equation 7 and the bandwidth is a known requirement, then equation 8 can also be used to solve for the Gain Constant after being set equal to unity at the Loop Bandwidth frequency:

$$\text{Gain Constant} = \frac{K_D K_V}{NCR_1} = \frac{(2\pi f_{BW})^2}{\left[1 + \left(\frac{f_{BW}}{f_{zero}}\right)^2\right]^{0.5}} \quad (9)$$

This leaves the parameters K_D, K_V, N, C, R_1, and R_2 to be determined. The parameters K_D and K_V are determined by component selection. N is determined by the frequency multiplication requirement. Thus, C, R_1 and R_2 remain to be calculated. From equation 7.

$$C = \frac{1}{2\pi R_2 f_{zero}} \quad (10)$$

Substitution of equation 10 into equation 9 results in the following resistance ratio:

$$\frac{R_2}{R_1} = \frac{N\,(\text{Gain Constant})}{K_D K_V\, 2\pi f_{zero}} \quad (11)$$

Unfortunately, there are two equations and three unknown variables. Therefore, engineering experience can be used to select component values. First select a reasonable capacitor value wnich will make the circuit component non-electrolytic or one with excellent frequency response characteristics. Now, equation 10 can be solved for R_2 and then R_1 can be calculated from equation 11. Check the resistance values calculated and compare these values to the range of resistances that the Phase Detector data sheet specifies. Also check the Op Amp specification sheet for gain (R_2/R_1) for frequencies less than the loop bandwidth and resistance ranges. If these critereon are not met then recalculate until they are achieved or find other devices which can accomodate the design calculations.

Examples

The effects of delay can be shown through examples. The PLL example design requires a 10 Hz Loop Bandwidth, a Damping Factor of 0.27, and a Reference Frequency of 100 KHz. Figure 2 shows a Bode Plot with the results of a Zero Frequency Location of 17 Hz and a Gain Constant of 71 dB. Note the Phase Margin is ≈ 30 degrees at 10 Hz and the magnitude is at 0 dB to give a 10 Hz Loop Bandwidth. These results are the common classic solutions seen in several references. Many references suggest no Loop Bandwidth limitations; however, unexplained bandwidth limitations were encountered in circuit applications. Note the reference frequency is a factor of 10,000 greater than the Loop Bandwidth.

A new plot can be calculated with the Reference Frequency at 100 Hz or 10 times greater than the Loop Bandwidth. The new plot is shown in Figure 3. One notices that the Phase Margin has gone negative and that results in an unstable loop. This demonstrates the dramatic effect of delay on loop performance.

A new solution for the example with reference frequency at 100 Hz can be calculated. Using the derived equations which include Divider Delay effects, results in Figure 4 which shows a 4 Hz Zero Frequency Location and a 64 dB Gain Constant and has a 30 degree Phase Margin (0.27 Damping Factor). Also notice the magnitude of 0 dB at 10 Hz (3dB BW = 15Hz from Equation 7.) results in a Loop Band width of 10 Hz. Thus an unstable loop in Figure 3 can be adjusted to meet loop requirements as shown in Figure 4.

With the establishment of these equations, a few tables can be used to aid PLL design. The first table in Figure 5 shows Damping Factor versus Loop Bandwidth to

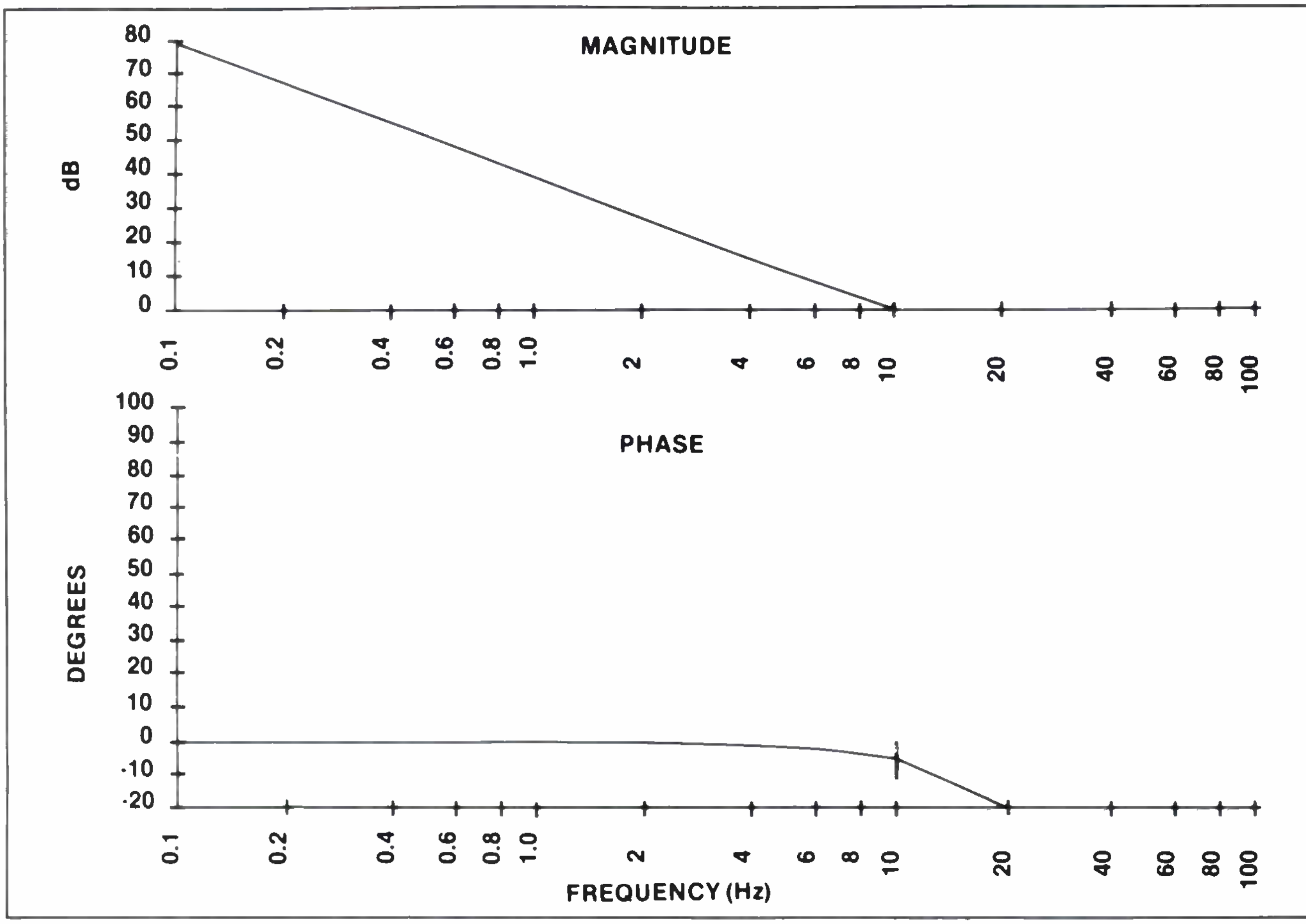

Figure 3. PLL open-loop transfer plot for reference freq- 100Hz. Zero freq location- 17Hz and gain constant- 71dB

DAMPING	BW/REF RATIO
0.10	0.22
0.20	0.19
0.30	0.16
0.40	0.13
0.50	0.11
0.60	0.09
0.70	0.07
0.80	0.06
0.90	0.05

Figure 5. Bandwidth to reference frequency ratio for a zero location at 0Hz.

DAMPING	ZERO/BW RATIO	GAIN CONST (dB 1Hz BW)
0.05	1.12	29.38
0.1	0.92	28.53
0.15	0.75	26.26
0.2	0.61	24.75
0.25	0.49	22.9
0.3	0.38	20.56
0.35	0.28	17.45
0.4	0.19	12.84
0.45	0.11	3.57
0.5	0.04	
0.55	negative	

Figure 6. Calculation of damping factor versus normalized zero location for BW/REF ratio of 0.1

Reference Frequency ratios. This table show ideally how close the Reference Frequency can be to achieve the Damping Factor shown. System choices in determining Loop Bandwidth, Reference Frequencies and stability can be aided with this table.

Figure 6 shows another table which locates the Zero Frequency Location relative to Loop Bandwidth for different Damping Factors at a 0.1 Loop Bandwidth to Reference Frequency ratio. Notice to achieve higher Damping Factor (greater stability), the Zero Frequency Location approaches 0 Hz, while the Gain Constant decreases to compensate for the Zero Frequency Location. This table gives insight into the Zero Frequency Locations effect on stability.

Conclusion

The results of this paper will allow more analysis of Phase Lock Loops before they are built and modifies the classic PLL approach to solve an increased number of PLL problems. It has been shown that for wide bandwidths the delay term is a dominanting factor in determining bandwidth. Thus, an aid to System Designers has been developed so that the feasibility of proposed Phase Lock Loops can be determined.

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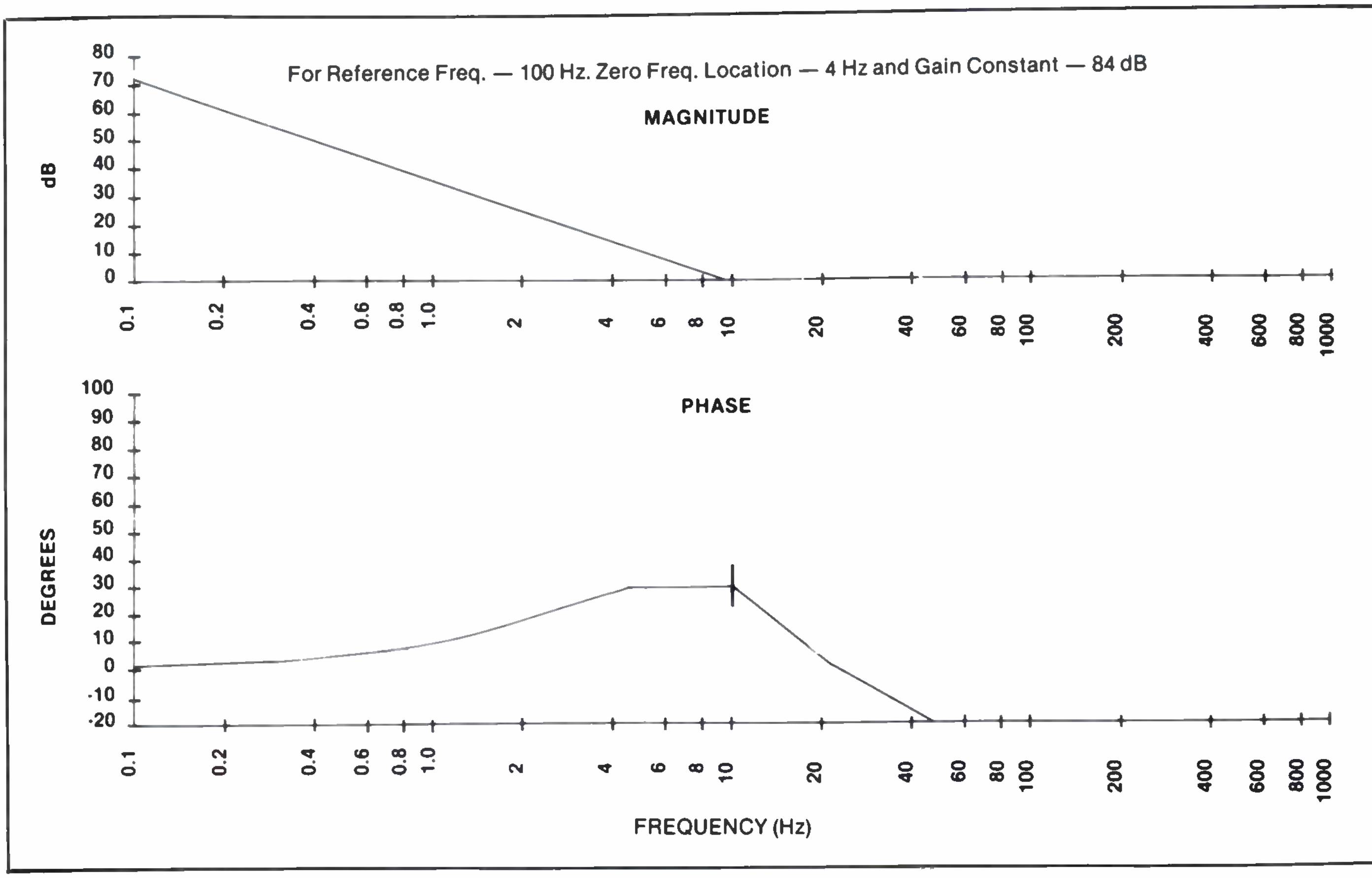

Figure 4. Modified PLL Solution.

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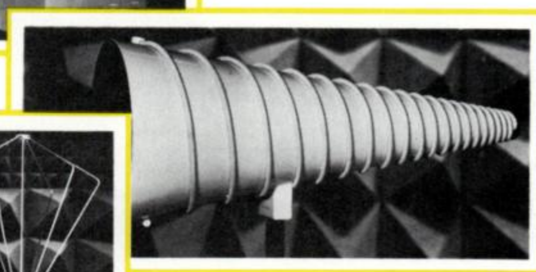

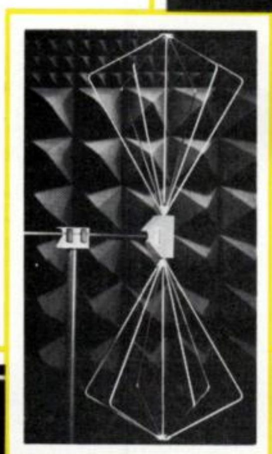

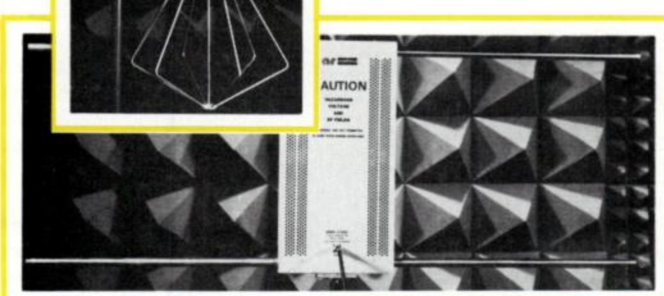

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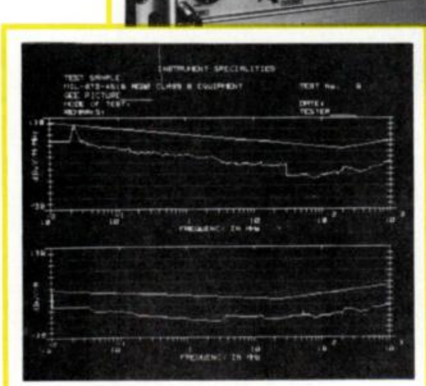

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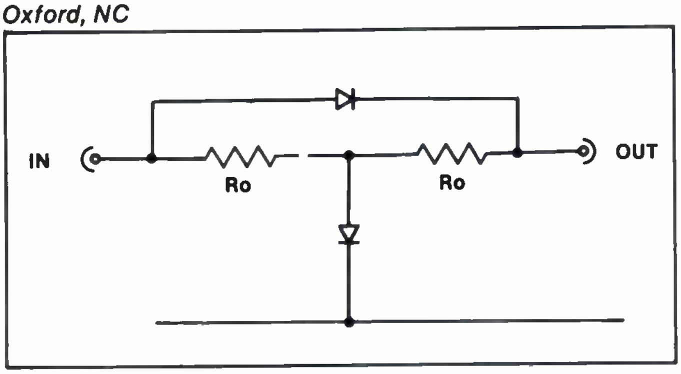
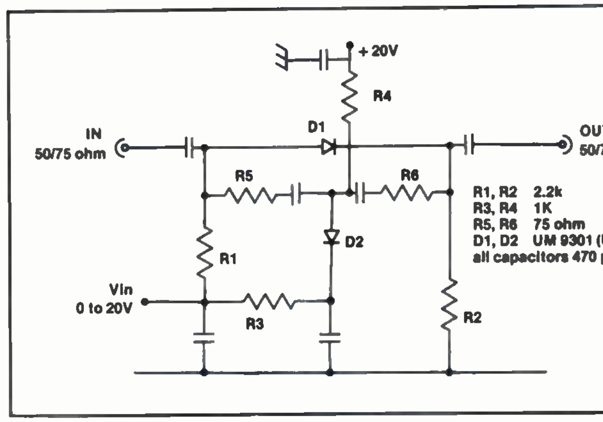


Bridged T Network



Actual Circuit Developed

Application

Broadband systems such as cable television a.g.c. circuits where intermodulation is of prime concern.

Circuit Description

Based on the well known bridged - T network shown above:

At high input voltage and low attenuation, D1 tends to conduct the signal. R1 and R2 set the current and isolate the DC. D2 tends to be off.

At low input voltage and high attenuation, D1 tends to be off. D2 tends to bypass the signal to ground. R3 and R4 set the current and isolate the DC. R5 and R6 maintain the characteristic impedance.

Performance -

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INPUT VOLTS	100 MHz LOSS dB	100 MHz R-LOSS dB	200 MHz LOSS dB	200 MHz R-LOSS dB	400 MHz LOSS dB	400 MHz R-LOS
20	2	18	2	24	3	[illegible]
16	4	26	4	20	5	[illegible]
12	6	23	6	17	6	[illegible]
8	8	28	8	18	8	[illegible]
4	10	24	10	20	10	[illegible]
2	14	18	14	18	13	[illegible]
1	18	15	18	15	18	[illegible]
0	22	14	21	14	18	[illegible]

Two Series Sections -

INPUT VOLTS	100 MHz LOSS dB	100 MHz R-LOSS dB	200 MHz LOSS dB	200 MHz R-LOSS dB	400 MH LOSS dB	400 MH R-L
20	4	15	4	16	5	[illegible]
18	7	23	7	18	8	[illegible]
14	10	16	10	14	10	[illegible]
8	15	20	15	15	14	[illegible]
5	19	30	18	18	16	[illegible]
3	24	25	23	19	20	[illegible]
2	29	20	28	18	24	[illegible]
1	38	15	35	15	28	[illegible]
0	46	13	40	14	31	[illegible]

application note

A 50-350 MHz, 50 OHM Amplifier With 24dB Gain

By Gary McCollum
California Eastern Laboratories, Inc.

Introduction

This two stage amplifier is designed to operate over the 50-350 MHz communication band. Using NEC's NE99532 transistor in a 50 ohm system, the design achieves noise figures at the band edges of 3.1 dB and 3.0 dB with a typical gain of +24 dB. P_{1dB} at the band edges is 11.6 dBm and 20.4 dBm. Construction is on a single sided copper clad G10 circuit board material.

Specifications

The electrical parameters of the NE99532 amplifier are shown below:

$V_{CC} = 18V, I_{CC} = 55mA$

Frequency (MHz)	NF (dB)	Gain (dB)	Input Return Loss (dB)	P_{1dB} (dBm)
50	3.1	+24.2	−20.0	11.6
100	2.6	+24.0	−17.0	13.6
150	2.2	+24.1	−22.0	16.5
200	2.3	+24.4	−13.5	19.7
250	2.4	+24.0	−24.0	21.0
300	2.6	+24.0	−21.5	21.8
350	3.0	+24.6	−20.0	20.4

Circuit Description

Figure 1 shows the amplifier's circuit design. The input circuit is a 50 ohm low pass/high pass network followed by a transformer for matching to the first NE99532 transistor. In the interstage matching network, inductor L_5 series tunes the output of the first stage transistor and is followed by a high pass filter to flatten the gain. The output circuit uses L_8 to series tune the output of the second stage transistor; the two-section high pass filter matches the output to 50 ohms. Any out-of-band response is stabilized by placing a 68 ohms resistor in parallel with an RF choke connected to the collector of each transistor. Figure 2 shows the printed circuit board's actual size and component layout.

Tuning And Performance

Tuning of the input, interstage and output circuits is accomplished by spreading the turns of the inductors to obtain the best matching conditions and frequency responses. Figures 3 and 4 show the performance of the prototype amplifier.

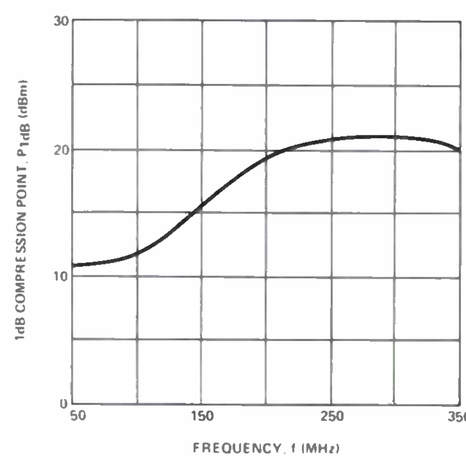


Figure 3. P_{1dB} vs. Frequency.

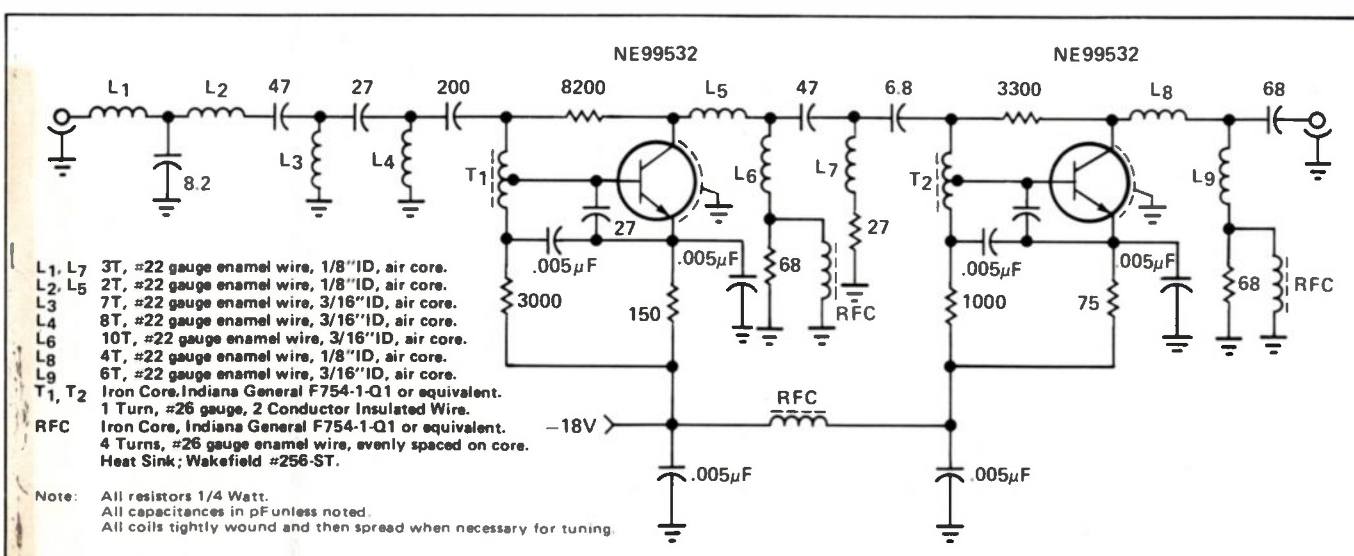


Figure 1. Amplifier Schematic.

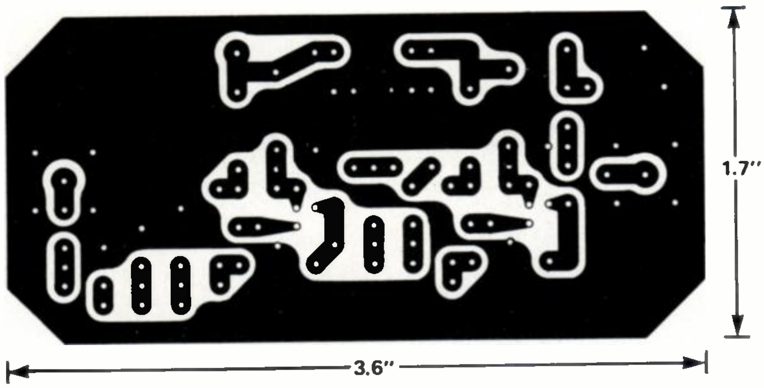
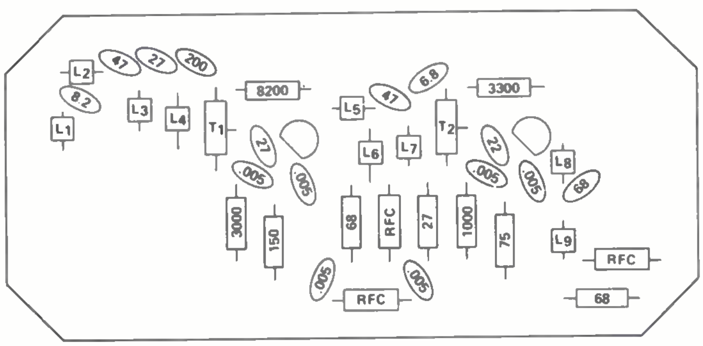


Figure 2. Printed Circuit Board and Layout; Actual Size.



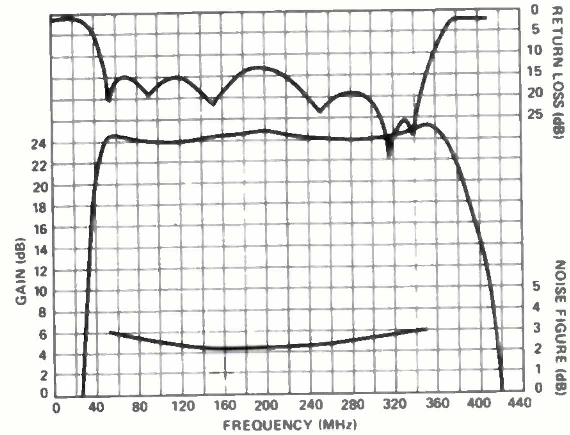


Figure 4. Gain/Return Loss/Noise Figure vs. Frequency.

Based on an application note, (AN83301) from California tern Laboratories, Inc.

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Wideband Microwave Network Analyzer

Hewlett-Packard Company's new HP 8510 network analyzer system brings to the microwave-design engineer what HP officials believe are truly profound microwave-measurement capabilities. Foremost among thes are:

- Error-corrected measurement of vector transmission and reflection parameters in real time, permitting real-time operator adjustments on the test device.
- Measurement accuracies that are 10 to 100 times more precise than previously attainable with commercially available instrumentation.
- 45-MHz to 26.5-GHz vector testing in a single setup with no change of connections.
- Optional transformation (using fast-Fourier-transform techniques) of error-corrected data between frequency and time domains at speeds permitting real-time adjustments.

The HP 8510 system can measure forward and reverse transmission and reflection characteristics (i.e., a network's scattering parameters) — measurements that have been vector-error corrected for systematic errors at 401 frequency points — in less than one second.

HP engineers believe that the HP 8510 system is the first commercial instrument able to measure magnitude and phase of transmission and reflection characteristics from 45 MHz to 26.5 GHz with a single test set. Furthermore, the measurement accuracies and resolution are believed to be 10 to 100 times better than is currently achievable in commercial practice.

List prices and current delivery estimates for the units comprising the new analyzer system are:

8510A, Network analyzer Option 010: Time-domain capability, $33,500 add $9,500, 12 weeks ARO; 8511A Frequency converter (45 MHz to 26.5 GHz), $17,300, 20 weeks ARO; 8512A Reflection/transmission test set (500 MHz to 18 GHz), $17,300, 12 weeks ARO; 8513A, Reflection/transmission test set (45 MHz to 26.5 GHz), $25,000, 20 weeks ARO; 8514A, S-parameter test set (500 MHz to 18 GHz), $28,000, 12 weeks ARO; and 8515A, S-parameter test set (45 MHz to 26.5 GHz), $37,200, 20 weeks ARO.

Hewlett-Packard Company, Palo Alto, CA 94304, Info/Card #110.

Spectrum Analyzer

As a leading supplier of portable and rugged spectrum analyzers, Tektronix, Inc. is adding a new set of "twins" to its 490 Series spectrum analyzer family. The two major new products — the 494 and its programmable compatible counterpart, the 494P — offer many features never before available in a spectrum analyzer. The new spectrum analyzers cover the frequency range from 10 KHz to 325 GHz and offer an exclusive "HELP" mode which provides explanations of controls and functions, on the CRT itself, at the push of a button. Fitting a competitive niche in the marketplace, the 494 and 494P will sell for $40,950 and $45,950 respectively. Delivery at introduction is eight weeks after receipt of customer orders. **Tektronix, Inc., Beaverton, OR 97077, Info/Card #109.**

L/C Meter

The Model 522 digital L/C meter and test fixture combination is being offered by Cambridge Technology, Inc at a new low price. This instrument combination allows measurement and sorting of inductances and capacitances at a basic accuracy of 0.25%, with dual measurement frequencies, and 0 to 10 VDC internal capacitor bias, for less than $1000; and industry first. A built-in comparator allows simultaneous sorting for high and low tolerances, and dissipation limits. Measurement frequencies of 120 Hz or 1 KHz are automatically selected. The instrument is autoranging and a range-hold feature permits fast repetitive measurements. The electrical contacts of the test fixture provide four terminal measurement capability. A push button allows both electrical contacts to be held open simultaneously so that components with very find leads can be inserted conveniently. A second

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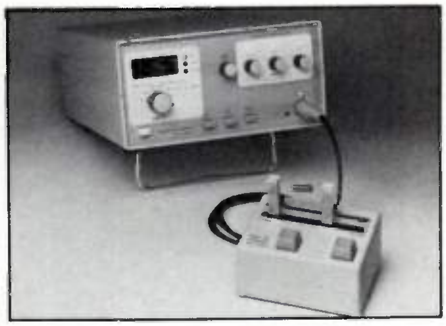


push button allows each set of contacts to be individually opened for components with flying leads such as coils or transformers. Other features include rugged all-metal enclosure, tilt stand, and lead capacitance null. The instrument is protected against damage from charged capcitors and comes with a one year warranty on materials and workmanship. A full refund within 30 days is available if the instrument is found unsatisfactory for any reason. The Model 522 is available from stock to 4 weeks and is priced at $875. **Cambridge Technology, Inc., Cambridge, MA 02140, Info/Card #107.**

Magnetic Field Strength Meter

Holaday Industries' new Isotropic Broadband Magnetic Field Strength Meter, Model HI-3006, measures magnetic fields from 300 KHz to 300 MHz. Our standard magnetic field probe covers the range from 5 MHz to 300 MHz and our new low frequency magnetic field probe covers the range from 300 KHz to 10 MHz. With this instrument, one can easily make accurate measurements of potentially hazardous fields in the vicinity of RF induction heaters, as well as near RF heat sealers, broadcast antennas, and other industrial and communications RF sources. The HI-3006 has all the features of the HI-3000 series including fully automatic zero adjust, audio output and peak reading memory. The HI-3006 comes complete with both probes, batteries and instruction manual in an attractive and protective carrying case. It is an ideal companion to the HI-3001 which, when configured with the optional high-sensitivity probe, can measure electric fields from 0.3 V/M to well over 3000 V/M. **Holaday Industries, Inc., Eden Prairie, MN 55344, Info/Card #106.**

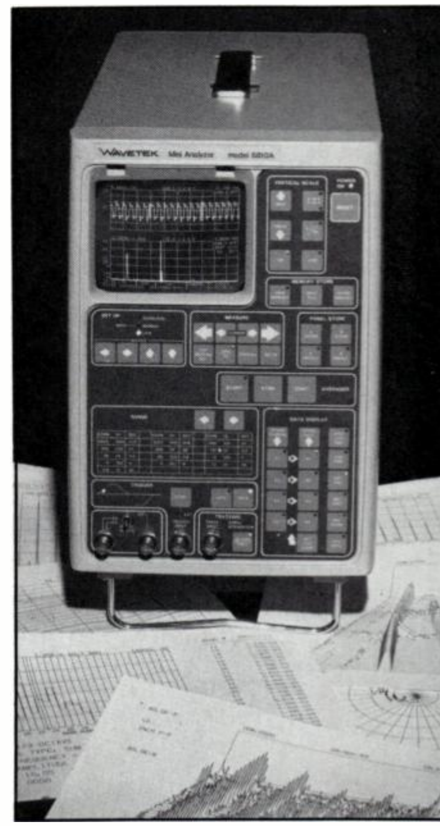
Frequency Synthesizer

With the introduction of two new synthesizers, PTS has created the most complete selection of cost-effective direct synthesizers. The PTS 500 now in production delivers low phase-noise and 0.1 Hz resolution at $7,300 while the PTS 040 has even

lower noise and equal resolution at $4,400. It addresses the ATE, Laboratory, and communications markets. Features common to all models of the line are: 200u sec switching, remote-only (OEM) models, choice of resolution, choice of integral interface BCD or GPIB. **Programmed Test Sources, Littleton, MA 01460, Info/Card #105.**

FFT Analyzer

Wavetek Rockland Scientific, Inc., has announced the introduction of their Model 5810A real-time spectrum analyzer, which incorporates extensive built-in data reduction to simplify the solution of typical measurement problems in vibration, acoustics, and electronics. The 400 line single channel analyzer also processes and displays new data at a high rate of 17 spectra per second, even while averaging as many as five spectra simultaneously (4 zoom and on baseband). Using a real time bandwidth of 7.2 KHz, the Model 5810A covers frequencies from 0.0025



Hz to 100,000 Hz. A built-in non volatile memory can store 200 spectra together with all information for later reconstruction of test conditions, including set-up, calibration, date, and time. Automatic storage at preset time intervals allows unattended logging of long term events such as drift or machine coastdown, or of short term data collection for stack ("waterfall") plots. Built-in data reduction routines permit data analysis in any of 11 different formats. These include stack plots of spectra at time or RPM increments, plots with the highest peak amplitudes and frequencies listed, time histories of four selected frequencies or orders of machine rotation, polar plots of phase vs. amplitude for balancing, and optional ⅓ octave and full octave conversions of live and stored narrow band spectra. Portable with on-line data storage, the Model 5810A is designed for use on-site. It also provides high sensitivity and wide coverage for laboratory testing. Simple set up, using any of 3 stored panels and 12 stored measurement cursor positions, insures repeatability in production or other high volume usage. In addition, the Model 5810A is GPIB compatible for fully automatic data collection and testing. Computational capabilities of the analyzer include single or double integration in time or frequency domains, PDS calculation, RMS overall or in spectrum regions, and consistent calibration in English or Metric units. Price of the Model 5810A is $12,400. Delivery is within 60 days after receipt of order. **Wavetek Rockland Scientific, Inc., Rockleigh, NJ 07647, Info/Card #103.**

Waveform/Vector Monitor Combination With SCH Capability

A rackmountable, single unit waveform/vector monitor combination from Tektronix, Inc., is now available for use by television personnel in all NTSC and PAL system countries. Similar in appearance and identical in mechanical configuration to the recently introduced 1740 Series, the new 1750 Series Waveform/Vector Monitor instruments offer a unique SCH phase (subscarrier to horizontal sync) monitoring and measurement capability, and full function vertical interval line

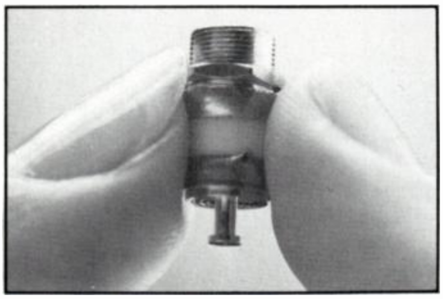
and field selection capabilities. The new 1750 (NTSC), 1751 (PAL), and 1752 (PAL-M) Series is intended to extend and compliment the line of television products offered by Tektronix. This new instrument provides unique SCH display capability, line and field select with LED readout, and the features of both the widely used 528A Waveform Monitor and 1420 Vectorscope Series. It requires the space of only one of the earlier units (5.25 inch, half rack) and costs just slightly more than a 528A/1420 combination. The 1750 (NTSC) and 1751 (APL) will each cost about $5,600. The 1752 (PAL-M) will cost about $6,160. The 1750 Series can be serviced at designated Tektronix facilities in the U.S.; and the company's service centers in Canada, Mexico, Japan, and Europe. Product availability is six weeks after the 1750's first public showing at the National Association of Broadcasters (NAB) 62nd Annual Convention and International Exposition, April 29-May 2, in Las Vegas, Nevada. **Tektronix, Inc., Beaverton, OR 97077, Info/Card #102.**

Hybrid Amplifier Module For Cellular Radio

A new series of three-stage common emitter hybrid amplifiers modules, designed and fully characterized for 800 MHz cellular mobile radio applications, has been introduced by Motorola. The new series (MHW808A1, MHW808A2 and MHW808A3) has a typical power output of 7.5 watts and a typical power gain up to 25 dB. Its class A input stage allows automatic gain control over 35 dB in output power range. The automatic gain control capability is necessary for cellular operation. The series employs three gain stages with transistors connected in a common-emitter configuration. The devices are specified at 12.5 volt, UHF characteristics. Included in the series are the following devices: MHW808A1, 806-870 MHz, 25 dB 7.5 W, $68.20; MHW-808A2, 806-890 MHz, 25 dB, 7.5 W, $74.80; and MHW808A3, 870-960 MHz, 23.2 dB, 7.5 W, $79.20. Availability is approximately 12 weeks from the factory and through authorized Motorola distributors. **Motorola Semiconductor Products Inc., Phoenix, AZ 85036, Info/Card #101.**

Air Variable Capacitors

Trim-Tronics, Inc., has introduced a line of air variable capacitors designed with a self resonant frequency greater than 5 GHz. The new multi-turn concentric ring capacitors complement Trim-Tronics tubular capacitor and air plate trimmer lines. They are suitable



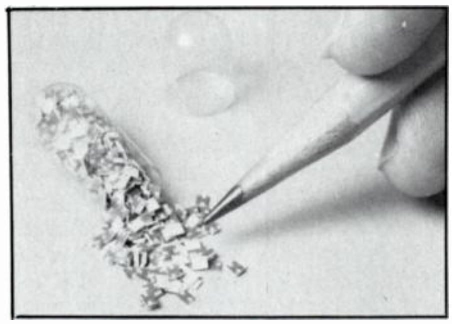
for sensitive telecommunications applications such as satellite, microwave, two-way radio and test instrumentation, where very precise tolerances are required. Typical functions include RF amplifiers, oscillators, crystal trimming, coupling, impedance matching and filter tuning. A unique component design produces a high Q-factor (greater than 5000 at 200 MHz), allowing the capacitor to operate at microwave frequencies. The capacitor's vertical slotted rotor mechanism results in complete surface area contact, producing uniform torque and contact resistance less than 1 milliohm. Available in several mounting styles, the Trim-Tronics air variable capacitor has a temperature coefficient of 0 ± 15 ppm over a temperature range from -55 to + 125 degrees C. it is rated to 250 V DC. **Trim-Tronics, Inc., Cazenovia, NY 13035, Info/Card #100.**

AVX Single-Layer Ceramic Capacitor Chip

A new single-layer ceramic and porcelain capacitor chip uniquely suited for stripline width matching in microwave integrated circuitry is now available from AVX. This new SLC is designed specifically for low-noise amplifiers, GaAs FET amplifiers, voltage-controlled oscillators, and other EW/ECM components. The new capacitor chip, called Pathguard™, employs sputtered electrodes with pure gold metallization and a leach-proof intermediate layer. The dielectric is fine-grained high-density ceramic with NPO, P090, X7R, or X7V temperature characteristics. Six different chip sizes are available depending on the capacitance required. The smallest chip measures nominally 0.015" x 0.015" x 0.006"; the largest measure nominally 0.090" x 0.090" x 0.010". These dimensions are precisely controlled for high repeatability. The capacitance range of the chips is from 0.5 pF to 2700 pF, with tolerances of ±0.1 pf to ±20 percent. Working voltage rating is at least 50 Vdc. Mechanical and environmental specifications are in accordance with applicable Mil Standards (883B and 202). Price of the chips ranges from $0.50 to $2.00 each in 1000-unit quantities. Delivery time is from stock to eight weeks. **AVX Corp., Myrtle Beach, SC 29577, Info/Card #99.**

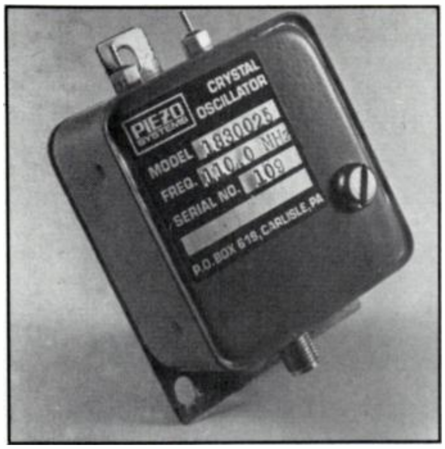
Minature Chip Attenuators

New miniature thin film chip attenuators from EMC Technology, Inc., are designed for high density circuit applications which also require stable, predictable performance. Model TS0500 chip attenuators are available in 1 to 20 dB values, with 3, 6 and 10 dB as standard, and for the frequency range from DC to 18 GHz. Attentuation accuracy is ±.25 dB. Measuring only .075" x .060" with .025" terminals, the new automatically trimmed attenuators require one-fourth the space used by similar attenuators. As a result, they are suitable for hybrid circuits of all kinds where packaging density of add-on devices as well as improved high frequency performance are important considerations. Attenuators may be solder mounted by all commonly used mehtods. EMC's new chip attenuators meet or exceed all applicable requirements of MIL-E-5400 and MIL-R-55342, and are fully RF tested through the X-band. For enhanced electrical performance, VSWR is 1.25 from DC through 4 GHz, 1.35 from 4-8 GHz, and 1.50 up to 18 GHz. Power dissipation is .1 watt, and impedance is 50 ohms. Attenuators operate from -55° to 150°C. Prices of the new Model TS0500 attenuators are based upon quantity and bandwidth desired; delivery is stock to 4 weeks A.R.O. **EMC Technology, Inc., Cherry Hill, NJ 08034, Info/Card #98.**



Crystal Oscillator

Piezo Systems has developed a crystal oscillator with improved SSB phase noise characteristics. The P/N 1830025 offered in the range of 50 to 110 MHz will provide SSB phase noise levels of -120 dbc/Hz at 100 Hz, -150 dbc/Hz at 1 kHz and -158 dbc/Hz at 10 kHz and beyond from the carrier, with a typical noise floor of -165 dbc/Hz (100 MHz measurements) the output power is 0 dbm into 50 ohms. (higher power available). The frequency stability is ±10 ppm over an operating temperature range of 0°C to +55°C. A volume of 3.3 cubic inches makes it attractive for compact systems where



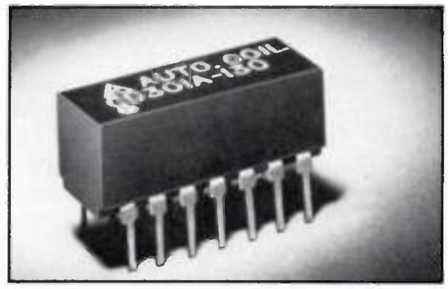
low phase noise is desired, such as synthesizers. The input voltage is + 15 VDC (others available). **Piezo Systems, Carlisle, PA 17013, Info/Card #96.**

20 GHz GaAs FET

Harris Microwave Semiconductor, a subsidiary of Harris Corporation, has announced the product release of its newest Gallium Arsenide Field Effect Transistor (GaAs FET), the HMF-0310. The HMF-0310 addresses the needs of broadband and narrow-band amplifier manufacturers for a high yield gain stage device which offers high gain and wide dynamic range at a reasonable cost. An extremely versatile GaAs FET, the HMF-0310 provides 15 dB gain at 8GHz, 8 dB at 18 GHz. Power bias and tuning conditions produces a typical linear power level of 100 mW (20 dBm) at 18 GHz. Low noise matching conditions typically result in a 1.2 dB Noise Figure at 8 GHz. Priced at $16.25 (US) each at 1000 pieces and designed for consistent performance and high yield through asembly, the HMF-0310 represents a cost-effective solution to a wide array of high frequency, high gain amplifier requirements. **Harris Microwave Semiconductor, Milpitas, CA 95035, Info/Card #95.**

Passive Delay Line

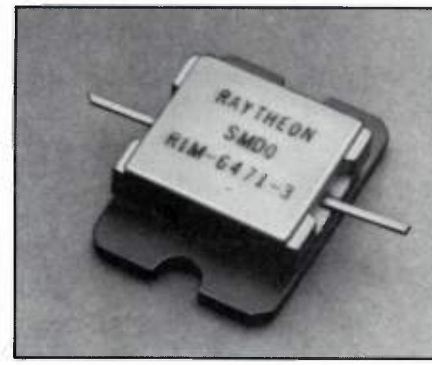
A new economy series of 14 pin DIP (Dual Inline Package) passive delay lines manufactured by Automatic Coil Corp. are compatible with various logic systems including TTL and video. These 180 standardized units, identified as the CD300 Series, have extensive applications in telecommunication, computer and instrumentation areas. These units are produced at the firms off-shore Carribean facilities to ensure extremely competitive pricing. They feature reliable hybrid construction and precise, stable delays in accordance with the specification of MIL-D-23859. Operating temperatures range from -55° to +125°C. Mounted dimensions of the units measure .800" long by .300" wide by .300" high. The solder coated phosphor bronze leads are .15" long and are evenly spaced at a center distance of .100". They are offered with overall delays ranging from 5 to 300 NS and with rise times ranging from 2.5 to 60.0 NS. Three basic styles are offered. One without taps and two styles with 10 taps each in different configurations. All units are available with standard impedances of 50, 100 or 200 ohms. Total delay tolerances are ± 2NS or 5% whichever is greater. Impedance tolerance is ± 10% and the attenuation standards are 10%



max. These units are also available to meet custom specifications which could include: intermediate delays, special taps or electrical parameters, tighter toleances, lower profiles, independent delay increments or internal terminations. **Automatic Coil Corp., Hialeah, FL 33010, Info/Card #93.**

Matched Power GaAs FETS

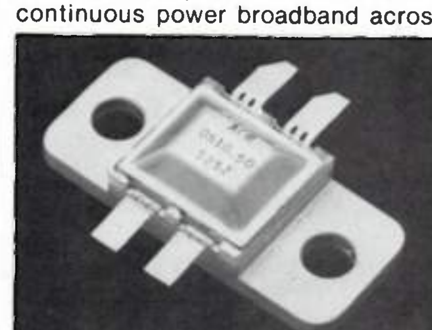
A series of internally-matched power GaAs FETs for operation at C-band has been introduced by Raytheon Company's Special Microwave Devices Operation. The RIM series consists of four devices that deliver 3 watts power at 3.7-4.2 GHz, 5.9-6.4 GHz, 6.4-7.1 GHz or 7.1-7.8 GHz. Each device contains a single 10 mm periphery pellet mounted between double-section matching networks. A hermetically-sealed, metal-ceramic package is utilized.



Raytheon's exclusive "via-hole" source connections makes possible low common-lead inductance. This results in highly-stable operation as well as virtual elimination of spurious oscillations. Other features include thinned chips with integral-plated heat sink for low thermal resistance, thick bonding pad metallization for bond strength, and dielectric passivation for environmental protection. RIM series internally-matched power GaAs FETs are priced at $300 each (1-9 quantity). Deliveries are 2-4 weeks after receipt of order. **Raytheon Co., Northborough, MA 02254, Info/Card #92.**

50-Watt Microwave Power Transistor

Acrian, Inc., has announced a 50-watt microwave power transistor for broadband amplifiers for military and commercial markets. The new transistor, called the Acrian 0510-50, is a Class AB part capable of supplying 50 watts (28 volts) of instantaneous and continuous power broadband across



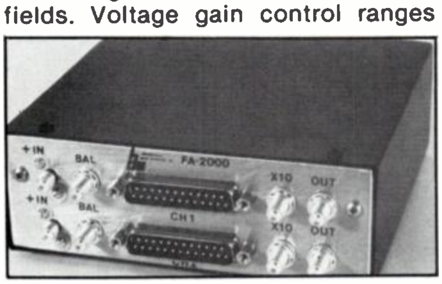
the 500-1000 MHz frequency range. The 0510-50 has high efficiency and good linearity. The 0510-50 has 50 percent collector efficiency and can withstand 5:1 VSWR. Maximum power dissipation is 125 watts, rated at 25 degrees centigrade case temperature. The 0510-50 is currently in full production and available for immediate shipment, at a cost of $210.80 in quantities of 1-99. **Acrian, Inc., Cupertino, CA 95014, Info/Card #141.**

VHF RF Amplifier

RF Gain, Ltd. introduces the RF6140V RF Power Amplifier in the 450-470 MHz band. With an input of 2-6 watts and output of 120-150 watts this 8 MHz spread amplifier features and anodized aluminum case heat sink. This makes the eighth model in their line of VHF amplifiers, including two repeater models. **RF Gain, Ltd, Rockville Centre, NY 11570, Info/Card #142.**

FA-2000 Filter Amplifier

HDS anounces the availability of the FA-2000 Filter Amplifier. The FA-2000 is a dual channel, low-noise, wide dynamic range video amplifier for use with low impedance sources. Remote control of gain, low frequency cutoff, and high frequency cutoff are provided. Connections for first stage output and external offset trimming are also provided. A nickel-iron enclosure shields against electric and magnetic fields. Voltage gain control ranges



from 10 to 10K; high pass filter control ranges from DC to 1 kHz; and low pass filter control ranges from 0.3 to 300 kHz. Features include 6 dB noise figure, 100 μV DC offset referenced to input, 50μA input bias current, and distortion less than 0.1% The FA-2000 is packaged in a 5.5 x 5.5 x 2.0 inch nickel-iron enclosure. Signal input and output connections are SMA. Control input is via a DB-25 connector. The FA-2000C Controller/Power Supply is available for remote operation. Single channel, fixed gain, and differential input versions are available, Delivery: Stock to 90 days. **HDS, Inc., Reston, VA 22091, Info/Card #143.**

Broadband Power Amplifier

A new generation of RF Amplifier, designed to replace those bulkier, less efficient and more costly tbe amplifiers. The new 3100LA is compact, lightweight (60 pounds) and capable of producing more power, over a wider frequency range than its predecessors. The 3100LA will operate from any ordinary single phase power receptable. The 3100LA will produce more than 100 Watts of linear Class A power and up to 180 watts of pulse and saturated CW power, with a useable operating range from 100 kHz to 180 MHz. The unit is unconditionally stable and will not oscillate for any condition of load and source impedance. Any signal generator, synthesizer or sweeper will supply adequate signal level to drive the 3100LA. The unit offers 55 dB of gain and will amplify AM, FM, SSB, TV and Pulse as well as other complex modulation with minimum distortion. Output RF Voltage level as well as power output into 50 ohms is monitored by a front panel meter. Both the integral power supply and forced air cooling system are conservatively designed to permit operations over a wide range of temperatures and AC line conditions. **Electronic Navigation Industries, Inc., Rochester, NY 14623-2881, Info/Card #144.**

Microwave Power Source

Antennas, waveguides and electronic components can now be tested and evaluated under microwave power using Cober's new Model 4169 Power Generator. Employing commercially available klystrons, the Model 4169 will deliver 2 kW of continuous power and can also be used in a pulse mode with duty factors to 100%. Power output and frequency of the Model 4169 is determined by the klystron selected. Typical tubes provide 2 kW at a tunable frequency in S band (1.7 to 2.4 GHz), C band (4.4 to 5.0 GHz) and X band (10.0 to 12.4 GHz). The Model 4169 will power other klystrons requiring beam supplies up to 10 kV at 1.25 A and 10V/10A floated filament supplies. Connecting a conventional PIN diode RF modulator to the unit's input, enables pulse mode operation. Pulse widths are narrow as 50 ns or repetition frequencies to 10 MHz are achievable. Under typical test conditions, 10 ms pulse widths at 10 KHz produce 200 W of average power. The Model 4169 is available as a single or dual tube configuration. In the latter, two klystrons are included in the cabinet permitting operator selection of a particular output of interest. A fully integrated and instrumented test console, the Model 4169 monitors key parameters with analog meters with overload relay protection. Safety interlocks are provided and equipment functions are displayed with clearly identified status lights. The Model 4169 is priced at $44,900 in the single tube configuration and $49,500 as a dual tube model. Klystrons are selected and provided by the customer. Delivery is 6 months. **Cober Electronics, Inc., Stamford, CT 06904, Info/Card #87.**

Yagi Antennas

Childs Corp has introduced two new control station Yagi antennas that are heavy-duty and yet are low in cost. The 6 element 806 to 870 MHz Yagi and the 3 element 144-174 MHz Yagi feature a square boom for rugged accurate element alignment together with heavy-duty 3/8 inch and/or 1/2 inch elements. The 800-Yagi comes factory tuned and covers the entire 806-870 MHz frequiency range with low VSWR. The Yagi-150 is field tuneable over the 144-174 MHz range for optimum low VSWR. All 3 models feature an extra large L-bracket mount that allows horizontal or vertical mounting to the tower, mast or wall mounting. The antennas are completely preassembled and tested at the factory for easy installation. **Childs Corp., Janesville, MN 56048, Info/Card #86.**

Dual Directional Coupler

Sage Laboratories, Inc., is now offering a compact, low-loss, dual directional coupler operating from 1-18 GHz.

Dimensions of this broadband coupler, model FC3428, are 3.70 x 8.50 x 0.6 inches. The unit has both a 10 dB coupler and a 15 dB coupler. Mainline insertion loss is 1.5 dB maximum, including coupling loss. Directivity is

15 dB minimum and the VSWR is 1.5/1 maximum. The coupler has a 1 dB roll off with frequency to compensate for system insertion loss-increase frequency. Delivery is 60 days. **Sage Laboratories, Inc., Natick, MA 01760, Info/Card #147.**

New Literature

Microwave Products Catalog

A new 44 page catalog covering attenuators and accessories for microwave applications has been published by Alan Industries, Inc. This edition updates electrical specifications for the company's programmable, cam actuated, rotary, manual switch and fixed attenuators as well as their accessories of loads, dividers, terminations, RF fuses and bridges. Among new attenuators in the catalog are a continuously variable series and additions to the programmable and manual switch series. New accessories include high frequency terminations and directional couplers. Product details are complete with photographs and line drawings. **Alan Industries, Columbus, IN 47202, Info/card #140.**

EMI/EMC Equipment Catalog

Tucker Electronics Company announces publication of a 52-page catalog featuring a full range of EMI and EMC equipment. This catalog, available now, contains descriptions and specifications on over 500 items. Included in the catalog are such items as: receivers — ELF, LF, VHF, microwave and surveillance; spectrum analyzers, antennas and probes, transient detectors and recorders, impulse and special purpose generators, power oscillators, signal sources, sweep generators, and power amplifiers. This equipment is manufactured by companies like Electro-Metrics, Singer-Stoddart, Empire, Collins, Honeywell, Hewlett-Packard and Tektronix. All equipment is expertly reconditioned and calibrated in Tucker's extensive lab facilities. A new feature of this

catalog is the Technical Book Store. This section contains a full selection of handbooks on EMC/EMI technology, including the six-volume library from Donald R.J. White. Another section is devoted to items new to Tucker's inventory. These include amplifiers, analyzers, bridges, frequency measuring, meters, microwave, scopes and probes, phase/synchro/resolver, recorders, and sweep and signal generators. All of these items are also reconditioned and calibrated. All reconditioned and calibrated equipment purchased from Tucker Electronics carries a 90-day warranty on parts and labor. In addition Tucker offers a 10-day return priviledge, with full refund, if the buyer is not completely satisfied. **Tucker Electronics Co., Garland, TX 75046, Please circle Info/card #138.**

Communications Catalog Cross-Reference Guide

RF Gain releases new first edition communications catalog/cross-reference guide. This new catalog contains never before published cross-references. In addition to expanded cross-references for Motorola, G.E., Johnson, RCA/TACTEC, Quintron, Regency, Wilson and Aerotron, and pricing at 20%-50% below the manufacturers dealer prices, the new catalog now includes cross-references for Standard, Repco and a comprehensive list of Japanese transistors for Midland, Yeasu, Force, Icom., etc. **Richardson Electronics, Ltd., Rockville Centre, NY 11570, Info/Card #139.**

Antenna Catalog

Watkins-Johnson Company has released a new 112-page catalog describing the company's entire line of antennas and antenna systems. The fully illustrated catalog details W-J's directional, omnidirectional, aerospace, direction-finding, search and surveillance, high-power and special-purpose antenna products. Compatible pedestals and controllers are also described in the catalog. **Watkins-Johnson Co., San Jose, CA 95131, Info/Card #137.**

Multi-Turn Capacitors Brochure

A new six-page brochure highlights the Trim-Tronics, Inc., line of air variable capacitors including the company's recently introduced multi-turn tubular capacitor. The 19 trimmers discussed are suitable for sensitive telecommunications applications such as satellite, microwave, two-way radio and test instrumentation where very precise tolerances are required, Capacitance of the line ranges from 3pF to 16pF, with four mounting styles to meet almost any application. The brochure includes photographs of each product, along with detailed dimensional diagrams, specifications and ordering information. **Trim-Tronics, Inc., Cazenovia, NY 13035, Info/Card #136.**

Military And Commercial Coil Catalog

A new 28-page catalog from Automatic Coil Corp. includes 655 military and commercial coils plus RF chokes. A highlight of the catalog, identified as AC-36, is a detailed 15-page coil specification guide for 5 popular military radios. These include the AN/PRC 104, AN/PRC 77, AN/VRC-12, AN/GRC-106 and WRC-1 radios. In addition to the Signal Corps module numbers used in each radio the catalog identifies all coils, toroids and filters used by Signal Corps part number and by Automatic Coil cross reference number. In addition, photographs of each radio plus all coil types are shown. The catalog will make it easier than ever for U.S. and overseas contractors to formulat their military radio component requirements. The commercial section of the catalog is devoted to standardized toroidal inductors and encapsulated toroidal RF chokes. Toroidal inductors are used in a myriad

DEMONSTRATION INFO/CARD 81
LITERATURE INFO/CARD 87

of applications including: EMI and RFI filters, power supplies, switching regulators, triac and SCR controls, transformers and loading coils. The catalog emphasizes that Automatic Coil has specialized in custom toroidal coil production for over 33 years because they represent a nearly perfect inductor. This is because their magnetic field is almost wholly confined within the core and the flux density is essentially uniform over the entire magnetic path. **Automatic Coil Corp., Hialeah, FL, 33010, Info/Card #135.**

Crystal Filter Catalog

Piezo Filters has introduced its new Crystal Filter catalog. Contained are technical specifications and data on standard crystal filters in the range of 2 to 155 MHz. A detailed description of the Engineering and Manufacturing capabilities of Piezo Filters is also included. Piezo Filters offers crystal filters in the range of 1 to 175 mHz with bandwidths of .01 to 3% of center frequency. **Piezo Filters, Carlisle, PA 17013, Info/Card #134.**

Power Inductors Data Sheet

The Inductive Products Division of TRW offers engineering specifications for its SR and LL lines of power inductors. The data sheet includes application, packaging, performance and military specifications. Included are charts listing inductance at 0 DC, maximum DCR, and current for each military part number and product type. Additionally, photographs, outline drawings and a chart provide size and construction information. SR and LL power inductors are recommended for switching regulator and AC filter choke applications. **Inductive Products Division TRW Electronic Components Group, New York, NY 10013, Info/Card #131.**

Cadec Software Package Brochure

Communications Consulting Corp. has just released a new brochure describing CADEC a computer aided design of electronic circuits software package. Available in standard and microwave versions, this software package is used for analysis and optimizations of any kind of electronic circuit on desk top computers in both the time and frequency domain. This 8-page brochure describes the features of CADEC, a program description, hardware requirements, technical data, and some typical circuit responses as analyzed by CADEC. **Communications Consulting Corp., Upper Saddle River, NJ 07458, Info/Card #130.**

Rectifier Bridges And Diodes Catalog

Electronic Devices, Inc.'s new catalog provides information on its line of standard, fast, and superfast recovery silicon rectifiers that includes silicon bridge rectifiers, high voltage axial lead rectifiers, high voltage packs and cartridges, and other rectifier devices. Extensive electrical and mechanical specifications are included, as well as information on new, state-of-the-art, 50ns. recovery rectifiers, 100 ampere rectifier bridges, and integral heat sink designs. **Electronic Devices, Inc., Yonkers, NY 10710, Info/Card #119.**

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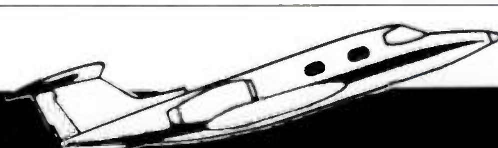

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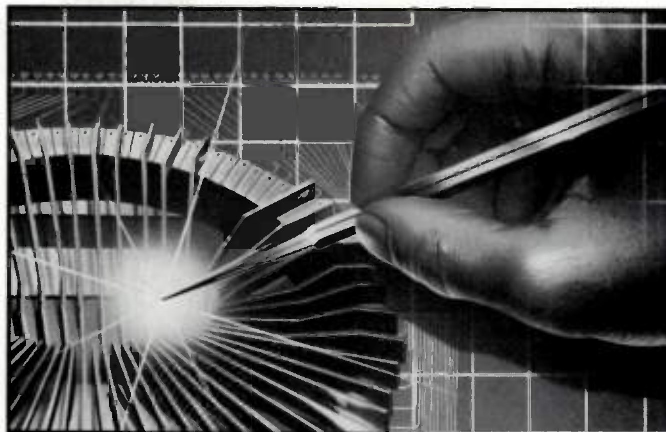

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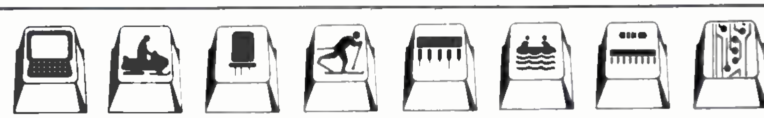


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