the GENERAL RADIO Experimenter

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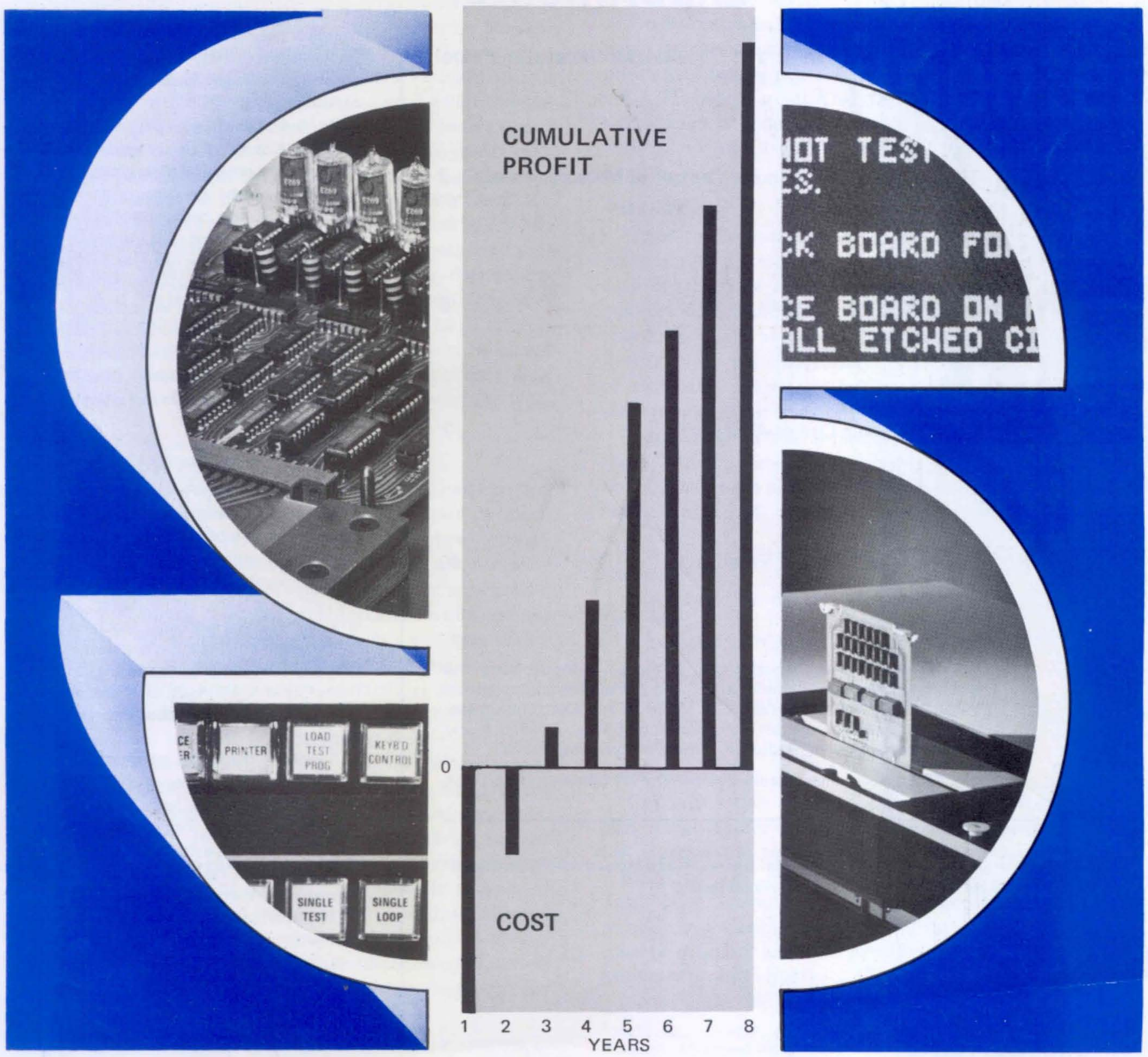

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The *General Radio Experimenter* is mailed without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

THE COVER should indicate to all readers that General Radio has proved the profitability of investing in a computer-controlled logic-circuit tester.

According to a forecast by the U.S. Department of Labor, issued early in 1970, the services and skills of engineers and technicians will continue to be in heavy demand during the next decade. The rate of growth of demand conceivably can be twice that of all working people.

The department foresees a shortage in supply of this critical group because of the decreasing enrollment in engineering studies. A determined effort must be made on the part of industry to upgrade all technical and semi-technical personnel to provide more efficient performance, thereby helping to close the gap between manpower needs and manpower accomplishments.

At the risk of stating the obvious, we would like to stress to our readers and their managers the need also to upgrade the use of machines and the ***machines*** themselves, releasing the human beings for more essential planning and thinking tasks. It is for this very reason that companies like General Radio recently have been devoting much of their research, innovation, and development talents to the production of automatic test equipment that will relieve technical personnel from repetitive and tedious tasks. This will permit more constructive and efficient use of their capabilities.

There is another basic reason for using machines: ***Profit.*** Industry finds that many applications of machine control to routine tasks save money, after the short period of time required to earn back (in most applications) the original capital investment in tooling. Too often, however, the men with the strongest instinct to do things better, faster, and more efficiently do not have sufficient background in finance to convince management of the ultimate wisdom of spending money, and sometimes lots of it, for machine assistance. It is to them we have addressed the first article in this issue.

Through the years of teaching fellow workers the ways to approach management for capital funds, we have found no better way than to present the proposal in terms of the probability of payback in a comparatively short time with realizable profits following immediately. We hope that you will be able to apply this principle to your own procurement problems.

C E White

C. E. White
Editor

The Economy of Computer-Controlled Measurements*

Introduction

Is your production schedule limited by manual tests of items completed or in process? Have you noticed that your inspection people are unable to retain efficiency as they routinely and monotonously check, check, check? Perhaps you've begun to give some thought toward changing your test methods, revising or replacing your old test equipment, and improving the efficiency of your inspectors. The idea of employing computer-assisted test equipment has been in your mind for some time but you don't know how to justify, to a hard-nosed management, the costs of the added facilities.

Your problem is no different from that which faced our production engineers at General Radio some time ago, when they decided to change from manual to computerized production-test operations. We thought, therefore, that a short discussion of the technique used by the engineers to convince GR management to finance the change would be of interest and value to our readers. The examples given use current cost and test-rate data and are presented for different quantities of digital logic boards to illustrate the application of the technique.

Why Economic Considerations?

The *engineering* decision to use costly test instrumentation is not very difficult since it is usually based upon technical considerations only. The *financial* decision, however, is usually made by an entirely different group of people, continually alert to the material needs and operating costs of an organization. Because of the ability of technical and financial minds to cross-fertilize each other and to reach a common understanding, progressive industrial organizations originated a rational approach to capitalization of facilities several years ago. They named it "cost-effectiveness" and it became a useful management tool. More important, it forced production engineers, in need of test equipment, to speak the language of accountants and broadened the appreciation each had for the other. Engineers began to speak in terms of total investment, discounted rate of return, depreciable life and tax shields. Accountants became equally appreciative of component failures and failure rates, of labor time to maintain equipment and to inspect production components, of equipment interfaces and software, and of precision of tolerances.

Another advantage of this cross-fertilization of ideas was to change the focus of management's attention from what *has been* invested in equipment to what *should be* invested in equipment. Yesterday's investment decision resulted in savings we are experiencing today, but can a new investment today result in even greater savings tomorrow?

These terms and techniques are illustrated below as they might be used to calculate a cost-effectiveness solution to the problem of testing digital logic boards. *Our* solution was the GR 1790 Logic-Circuit Analyzer.[1] The typical data used are based upon experience with GR logic circuits constructed with 12 to 60 14- or 16-pin digital IC's on printed-circuit boards.

The Old Way

Prior to the installation of the GR 1790 system in our manufacturing facility, we used hard-wired test fixtures for each board to be tested. Preparation of these fixtures required a design and fabrication period of 1 to 4 weeks per board, normally averaging 2 weeks. Test times using these fixtures were reasonably short (5 to 10 minutes), but the lack of significant diagnostic information resulted in troubleshooting and repair times of 20 to 40 minutes.

Total costs for this approach are shown below, assuming the minimum times given above, a quantity of 10,000-boards/year made up of 50 different board types, and a board failure rate of 33%.

WITH ORIGINAL TEST FIXTURES

Preparation:	
(50 types/yr) (2 wk/type) (40 h/wk) ($4/h)	$16,000
Test:	
(10,000 bd/yr) (5 min/bd) ($\frac{1}{60}$ h/min) ($4/h)	3,333
Troubleshooting and Repair:	
(3,333 rejects/yr) (20 min/bd) ($\frac{1}{60}$ h/min) ($4/hr)	4,444
Total	$23,777

The Forecast

Our production engineers estimated that, after introduction of a computer-controlled test system, preparation of the test programs and test fixtures would take 24 hours per board type. Actual test time per board, by relatively unskilled labor, would be 30 seconds. Since rejected boards would be accompanied by a troubleshooting printout from the computer, time to diagnose and to repair the rejects was expected to decrease from 20 minutes to 12 minutes.

*As applied to procurement and application of the GR 1790 Logic-Circuit Analyzer, described on page 7.

[1] Fichtenbaum, M. L., "Computer-Controlled Testing Can Be Fast and Reliable and Economical without Extensive Operator Training," *Electronics*, January 19, 1970.

Based upon these estimates, costs (in 1968) were calculated:

WITH COMPUTER-CONTROLLED TEST SYSTEM

Preparation:	
(50 types/yr) (24 h/type) ($4.00/h)	$4,800
Test:	
(10,000 bd/yr) (30 s/bd) ($\frac{1}{60 \times 60}$ h/s) ($1.65/h)	138
Troubleshooting and Repair:	
(10,000 X .33 bd/yr) (12 min/bd) ($\frac{1}{60}$ h/min) ($4.00/h)	2,640
Total	$7,578

Hence the total direct-labor savings made possible by use of the computer-controlled test system were estimated to be $23,777 – $7,578 = $16,199/year. Management approved the installation after reviewing these figures and studying a funds-flow analysis similar to that of Tables 3 and 4.

The New Era

Use of the 1790 Logic-Circuit Analyzers in the manufacturing facilities confirmed the production engineers' forecast. The preparation time was significantly reduced, since only a simple mechanical interface and an easy-to-write test program were required for each new device. These are normally prepared in ½-2 days depending upon the complexity of the board to be tested and, in our experience, averaged 1 day/type. The actual test time was reduced to milliseconds, but the time required to insert and remove the board being tested kept the total test time at an average of 30 seconds. The GR 1790 makes convenient the inclusion of diagnostic suggestions in the test program so that troubleshooting time may also be reduced. The time required, however, to effect a repair (replace an IC, remove a solder bridge, etc) kept the troubleshooting/repair time to an average of 6 minutes. Actual total costs for the same quantities used in the forecast to management are

WITH THE GR 1790 LOGIC-CIRCUIT ANALYZER

Preparation:	
(50 types/yr) (1 d/type) (8 h/d) ($4/h)	$1,600
Test:	
(10,000 bd/yr) (30 s/bd) (1/3600 h/s) ($2/h*)	167
Troubleshooting and Repair:	
(3,333 rejects/yr) (6 min/reject) ($\frac{1}{60}$ h/min) ($4/h)	1,333
Total	$3,100

*Relatively unskilled labor cost – 1969.

Hence, the total direct-labor savings made possible by the GR 1790 in this example are $23,777 – $3,100 = $20,677/year.

The typical quantities (and hence the labor savings) will obviously differ with industry and product. Table 1 gives the value of annual labor savings for 3 quantities of boards, 3 numbers of different types of boards, and 2 failure percentages. These figures are based upon the same rates used in the preceding example.

The saving in labor costs is only one calculation in the cost-effectiveness approach. It is also important to consider the expenses and savings over a period of time of concern (the cash flow) and to discount future funds to reflect their present value.**

The obvious initial expense is the purchase price of the system. Additional costs include time spent attending training courses and acquiring proficiency in writing test programs and using the system, plus normal operation and maintenance costs.

The labor savings calculated above are reduced by the 50% Federal corporate tax rate, as are other internal expenses and savings. Included on the savings side of the ledger is depreciation, a non-cash expense that acts as a tax shield. Analysis of the depreciation of the GR 1790 appears in Table 2.

Table 3 gives the funds-flow analysis for an eight-year period. The Net Operating Advantage is shown at the bottom of each column. Table 4 presents an analysis of the funds-flow after taxes for the same eight-year period. It is obvious from Table 1 that use of a larger number of different types of boards or a larger quantity of boards would significantly affect the final calculation. For example, if this study had been based on 100 different types of boards instead of 50, the Payback Period would have been about 8 months and the Discounted Rate of Return would have been approximately 150%.

**The application of accounting principles, which reflects the time value of money.

Table 1
Typical Annual Labor Savings
(Based on GR experience)

	Number of different board types					
	10% Reject			50% Reject		
Bd/yr	50	100	500	50	100	500
1,000	$14,800	$29,200	$124,400	$15,200	$ 29,600	$144,800
10,000	18,500	32,900	148,100	22,200	37,600	151,900
100,000	55,400	69,800	185,000	92,700	107,100	222,300

Table 2
Depreciation Calculation for GR 1790
(Sum-of-the-years-digits method)

Original cost: $32,500
Salvage: –4,000
Depreciable cost: $28,500

Useful life: 8 years

Year	Digits	Depreciation	50% Tax Shield
1969	8/36	$ 6,300	$ 3,150
1970	7/36	5,500	2,750
1971	6/36	4,700	2,350
1972	5/36	4,000	2,000
1973	4/36	3,200	1,600
1974	3/36	2,400	1,200
1975	2/36	1,600	800
1976	1/36	800	400
		$28,500	$14,250

Table 3
Funds-Flow Analysis — Type 1790

	1969	1970	1971	1972	1973	1974	1975	1976
EXPENSES								
Cash Outlay (Purchase)	$32,500	$ —	$ —	$ —	$ —	$ —	$ —	$ —
Cash Inflow (Salvage)	—	—	—	—	—	—	—	(4,000)
Production Engineering $1000 first year, $500 thereafter (50% tax shield)	500	250	250	250	250	250	250	250
Maintenance	500	500	500	500	500	500	500	500
Total Expenses	33,500	750	750	750	750	750	750	(3,250)
SAVINGS								
Test/Repair Labor Savings (50% Tax Shield)	10,338	10,338	10,338	10,338	10,338	10,338	10,338	10,338
Depreciation (50% Tax Shield from Table 2)	3,150	2,750	2,350	2,250	1,600	1,200	800	400
Total Savings	13,488	13,088	12,688	12,338	11,938	11,538	11,138	10,738
NET OPERATING ADVANTAGE	**($20,012)**	**$12,338**	**$11,938**	**$11,588**	**$11,188**	**$10,788**	**$10,388**	**$13,988**

Table 4
Funds-Flow After Taxes

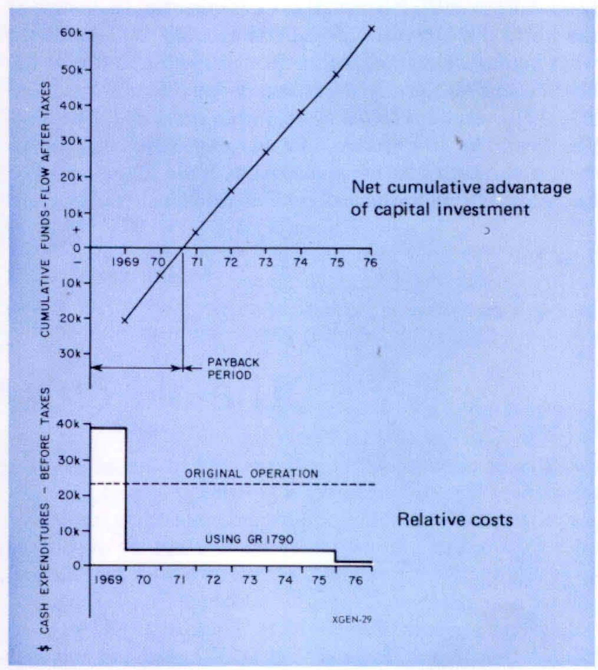
Year	Annual	Cumulative	Payback Ratio
1969	($20,012)	($20,012)	—
1970	12,338	(7,674)	—
1971	11,938	4,264	0.13
1972	11,588	15,852	0.49
1973	11,188	27,040	0.83
1974	10,788	37,828	1.16
1975	10,388	48,216	1.48
1976	13,988	62,204	1.91

Payback Period = 2.6 years

Discounted Rate of Return = 56%

A standard discount table was used to calculate the Discounted Rate of Return, which was 56%. This percentage can be related to a corporate goal for return on investment to screen out undesirable projects or programs.

Alternatively, an arbitrarily chosen discount rate, which approximates the desired internal rate of return, can be used to discount the cash flows. The Present Value of a project is the sum of the discounted cash flows; a positive Present Value indicates a profitable project. The magnitude of Present Value of a project can be related to that of other projects to allow management to make a choice between programs competing for available funds.



For those of you who are not familiar with cash flow discounting we offer a short explanation. We have referred to the time value of money. Because of this factor, expenses (cash outflows) of one period of time cannot be directly compared with income (cash inflows) of another period. The reason for this is that the money we have today can be invested to bring us a return and, therefore, will be worth more at the end of this year, next year, and each succeeding year that the money remains invested. At a discount rate of 10%, $1 earned three years from now is worth, to us today, $1/$1.33 or $.75. *That* is the Present Value of the $1 earned three years hence: $.75. The discount rate chosen is usually the desired *internal rate of return*.

Further information is available to readers interested in financial aspects of facility acquirement in a reprint of a talk by W. D. Hill of General Radio to the Planning Executives Institute, October 4, 1968, entitled "Planning Investments in Research and Development."

Other Applications

At General Radio, the GR 1790 is also used in Incoming Inspection for functional tests of all digital IC's. This inspection has reduced our failure rate of IC's in printed circuit boards from an initial 4-8% to less than 1%. Were these figures included in the cost-effectiveness analysis, the case for the GR 1790 would be even stronger. We did not, however, include these figures in the above example since the primary purpose of the GR 1790 is to test and troubleshoot assembled logic boards, and because relatively low-cost digital IC testers are available. On the other hand, the increasing use of MSI and LSI circuits in standard 16- and 24-pin packages has created additional testing requirements that cannot be met by low-cost digital IC testers. The ease with which the GR 1790 makes these tests assures ready customer acceptance even in its IC testing role. And, of course, printed circuit boards that use many of these MSI packages are in turn so much more complex that the reduction in test and troubleshooting time provided by the GR 1790 far exceeds the savings depicted in the examples above.

Views of the Manufacturing Manager

The planning and foresight of the production engineers were justified on the basis of simple dollars-and-cents analyses, before and after the fact. Consequently, their view of the world through rose-colored glasses could be excused. But what about the manufacturing manager, close to the assembly line and continually alert to every-day personnel relationships? His reactions to the system were expressed somewhat like this: **The test system, like any expensive tool, had to meet a number of basic requirements. It did. These included ease of operation by normally skilled machinists/technicians. The system was completely useful almost from the moment of installation – familiarization/training time was a minimum. The test capability of the system was broad, sufficient to permit change of interface equipment from component testing to assembly testing within a very short period of time. Vendor service, such as programming advice or advice on instrumentation implementation, was continually available from knowledgeable sources.**

The position of the manufacturing manager at GR is not necessarily the same as that of a manufacturing manager at another company. In this case, however, a true vendor-customer relationship existed because of the complexity of design and application problems. Consequently, the solutions to the personnel-interface problems between manufacturing and engineering were worked out smoothly and, in fact, became the basis for the program of service decided upon to implement the sales of the system to industry at large.

Conclusion

In many ways, our experience in development and application of the GR 1790 supports the theme that innovative metrology is, in fact, the key to industrial progress.* Industry can gain immeasurably by new ways of saving time and reducing costs and by new technologies and their applications.

*The theme of the 1970 Standards Laboratory Conference, sponsored by the National Conference of Standards Laboratories, is "Innovative Metrology – Key to Progress."

The Editor is indebted for most of the material contained in this article to P. H. Goebel, R. E. Anderson, and R. F. DeBoalt. Financing details were verified and expanded upon by W. D. Hill.

"The old order changeth . . ."

As companies grow, old patterns tend to change. Our International Division is currently growing at a rapid pace; we are progressively assuming a more and more direct role in our sales abroad, and old marketing relationships are dissolving.

In Europe we are establishing our own sales subsidiaries, and we have taken over from old and valued friends the job of selling and servicing GR products. Thus, in 1964, we established General Radio Company (U.K.), Limited and said good-bye to Claude Lyons, Limited after 27 years. As of the middle of 1969 we purchased the GR segment of Etablissements Radiophon, our French outlet for over 33 years, and rechristened it General Radio France, with Paul Fabricant temporarily staying on as President to ease the transition.

In setting up our new subsidiary, General Radio Italia S.p.A., and bidding farewell to Ing. S. and Dr. Guido Belotti S.r.l., we again bring to a conclusion a long and fruitful collaboration. Dr. Belotti, and his father before him, represented us in Italy for 37 years and will continue to manufacture, under GR license, Variac® autotransformers.

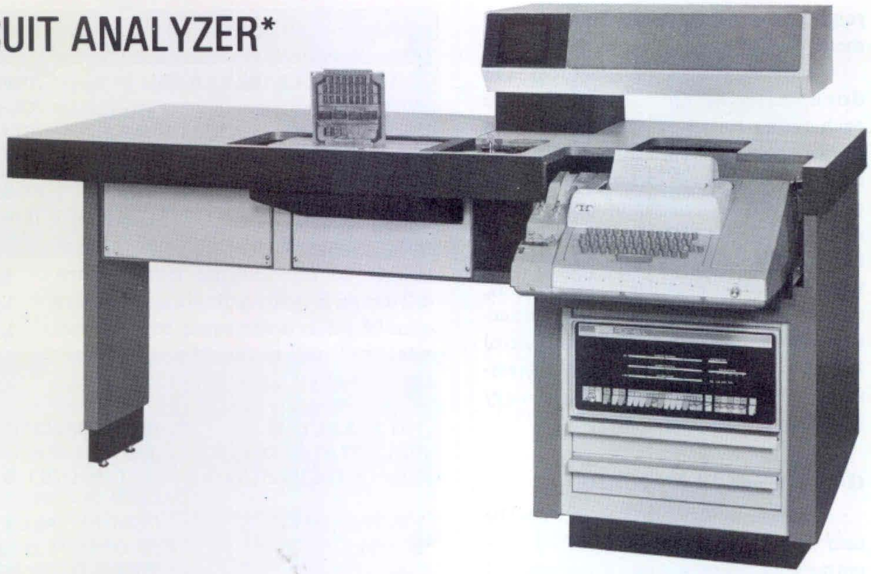
We have expanded the coverage of our German subsidiary, General Radio GmbH, to the northern part of Germany as well as the southern. This brought to a close a shorter association with Dr.-Ing. G. Nüsslein but one that has helped significantly to expand GR's market in Germany.

In Latin America we have worked for 29 years through the export house of Ad. Auriema, Inc. In furthering our objective of establishing as direct contact as possible with our customers, we are now moving one step closer to them by replacing this channel by a network of representatives directly responsible to GR. To Carlos Auriema, who, with his father before him, has been our colleague and friend, we must now say goodbye.

These gentlemen – Claude Lyons, Paul Fabricant, Dr. Guido Belotti, Dr. Günter Nüsslein, and Carlos Auriema – have all been good friends, as well as business associates, of GR. We wish them well in their continuing pursuits and thank them for their contributions to General Radio's successes.

– D.B. Sinclair

GR 1790 LOGIC-CIRCUIT ANALYZER*



GR 1790 DEFINED

The GR 1790 Logic-Circuit Analyzer is a computer-controlled functional GO/NO-GO and diagnostic test system for logic devices ranging from basic 14-pin integrated circuits to assemblies with as many as 96 inputs and 144 outputs. The system performs up to 4000 tests per second and yields a GO/NO-GO indication and a typewritten or scope-displayed error message.

Purchase justification is easy.** Savings are stressed in the process of programming and in the ready adaptability of test fixtures. Test programs are written by technician-level personnel in much less time than it takes to write manual test procedures, and costly tooling is eliminated by the simple and flexible device adaptor between the tester and the tested. Testing costs are low because of the speed of computer-directed tests, and troubleshooting costs can be sharply reduced by the inclusion of operator diagnostic instructions in the test program.

The simplified test language developed by General Radio can be learned in just a few hours. The entire test operation is characterized by speed and efficiency:

1. The operator writes a test program consisting of simple statements of the input and output conditions of the circuit to be tested.
2. The test program is converted to punched tape on the teletypewriter and then is automatically translated into a more compact form; programming errors are detected during the translation.
3. The test program is entered into the computer via the high-speed tape reader.
4. The test circuit is connected to the system by a device adaptor corresponding to the input/output configuration of the circuit.
5. The operator presses the START button on the control panel.

All testing then proceeds automatically. Should a fault occur, the operator can troubleshoot immediately or continue to test the remainder of the devices, saving the repair work for later.

The five steps above apply only when a new test program is required. If the test operator receives a new manufacturing run of a previously tested device only steps 3, 4, and 5 will be needed, thereby enhancing the speed and savings features of the GR 1790.

THE PHYSICAL ORDER

The standard system components of the GR 1790 are:

- Computer with 4,096 12-bit words of 1.6-μs-cycle core memory.
- Interface system.
- Control panel.
- Power supplies.
- Teletypewriter with keyboard, tape reader, and tape punch.
- Photoelectric tape reader.
- Alpha-numeric display oscilloscope.
- Logic probe.
- Device adaptor kits.

Options include a rack version, additional memory, and programmable logic levels.

In both desk console and rack versions, all controls are within easy reach and monitoring indicators are readily visible (Figure 1). The GR 1790 does not

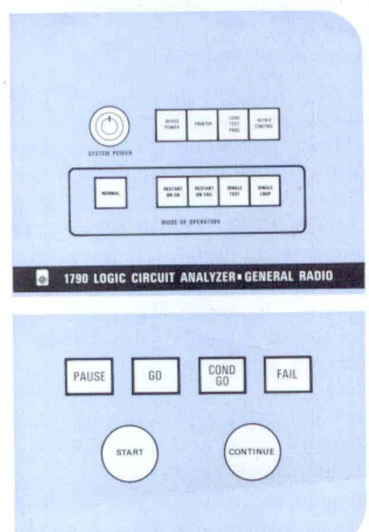


Figure 1. Control panel layout.

*Abstracted from special brochure available upon request.
**See page 3.

require a special, controlled environment.

Costly tooling, test fixtures, and documentation are eliminated since each circuit to be tested requires only a device adaptor and a test program. The device adaptor accommodates most physical configurations of test circuits and is locked in place in a recess in the top of the console. Adaptors, provided with wire-wrap pins on etched boards, can be wired with a variety of connectors and can include additional control functions, loads, logic-level conversions, or any other circuitry necessary for the application at hand.

GR 1790 TEST LANGUAGE

A high-level test language enables the user of the analyzer to write test programs without requiring a knowledge of computer programming. This language is described in detail in the special brochure. A few of the many features of this test language are noted below:

Autoprogramming

It is not necessary to program the output logic states of the device to be tested. The GR 1790's Autoprogramming feature enables a known good reference device to provide this information to the system. After automatically double-checking these outputs (against a second known good reference device, if desired), the computer stores them as a permanent part of the test program. Hence, no reference device need be kept on hand during the actual testing.

Automatic Generation of Tests

By definition, only completely combinational logic may be tested with an arbitrary pattern of input stimuli. Such logic circuitry may require as many as 2^n tests for a device with "n" inputs. The GR 1790 SEQUENCE statement eliminates all the effort in test programming by automatically generating a sequence of tests with all combinations of the specified inputs. The outputs of a "known good" reference device are stored by the Autoprogramming feature described above. Use of the SEQUENCE statement results in an extremely simple program (A).

Diagnostics

The GR 1790 test language facilitates inclusion of diagnostic informa-

A

```
*I(1,2,3,5,11,4,7,10,12,6,9,8)            /INPUT SPECIFICATION STATEMENT
*0(43,44,61,69,82)                        /OUTPUT SPECIFICATION STATEMENT
SEQUENCE(3,5,1,2,7,12,11,10,4,8,6,9)      /STATEMENT WHICH AUTOMATICALLY
                                          /GENERATES A SEQUENCE OF 4,096
                                          /TESTS WITH ALL COMBINATIONS OF
                                          /THE LISTED INPUTS
END                                       /END OF PROGRAM
```

B

```
*I(1,13,5,81,7,19)          /INPUT SPECIFICATION STATEMENT
*0(1,2,37,62,69,71,49,50)   /OUTPUT SPECIFICATION STATEMENT
1;IH(1,5,7)IL(13,81,19)$    /SET INPUTS 1, 5, AND 7 HIGH AND 13, 81,
                            /AND 19 LOW; DON'T CHECK OUTPUTS ($)
IGNORE(#37,62)              /IGNORE "ALL BUT" (#) OUTPUTS 37 AND 62
IF(37)2                     /IF OUTPUT 37 IS HIGH AND 62 IS LOW (DESIRED
                            /RESULT) TRANSFER TO TEST 2
PRINT CHANGE IC 14!         /INCORRECT RESULT, SO DISPLAY A MESSAGE
                            /TO OPERATOR ON SCOPE
PAUSE 1                     /SYSTEM PAUSES
2;IL(1)OL(#)                /LOWER INPUT 1 AND TEST THAT ALL OUTPUTS
                            /ARE LOW
.
.
.
33;IH(13,5)IL(#13,5)$       /SET INPUTS 13 AND 5 HIGH AND ALL BUT 13
                            /AND 5 LOW
IGNORE (#69,71,49,50,2)     /IGNORE ALL BUT THESE OUTPUTS
IF (#69)34                  /IF ALL BUT OUTPUT 69 ARE HIGH AND 69 IS
                            /LOW, TRANSFER TO TEST 34
CALL 70                     /CALL DIAGNOSTIC SUBROUTINE BEGINNING
                            /ON TEST 70 AND RETURN HERE
34;                         /NEXT TEST
.
.
.
DO 53,100                   /DO LOOP WHICH REPEATS THE NEXT TEST
                            /THROUGH TEST 53 ONE HUNDRED TIMES
50;IH(1)$                   /DESIRED SEQUENCE OF INPUTS TO BE REPEATED
51;IH(13)$
52;IL(1)$
53;IL(13)$
.
.
.
70;$                        /DUMMY TEST – BEGINNING OF SUBROUTINE
PRINT ATTACH IC CLIP        /DIAGNOSTIC ROUTINE USES OPERATOR
   TO IC34, THEN            /INTERVENTION
   PRESS CONTINUE!
PAUSE 59                    /SYSTEM PAUSES
71;                         /DIAGNOSTIC ROUTINE
.
.
78;
RETURN                      /PROGRAM RETURNS TO LOCATION FROM WHICH
                            /IT WAS CALLED
END
```

tion in the test program. When failures occur, the program can transfer to diagnostic routines or display messages to the operator suggesting possible remedies. Some examples of diagnostic tests and other test-language features are contained in sample program B.

THE END RESULT

Tests of circuit boards at GR's West Concord facility serve as examples of applications of the features described on the preceding pages. The fact that these boards (and many others like them) are now being tested in volume on the GR 1790 and incorporated into other GR instruments is testimony to the speed, economy, and effectiveness of the Logic-Circuit Analyzer.

The first sample board, Figure 2, with 10 inputs and 6 outputs, consists of 40 integrated circuits and 22 discrete components (functionally, two 12-bit binary counters, one 24-bit recognition circuit, and six state-recognition circuits). The programming time required for this board was 12 hours. Device-adaptor preparation involved only wiring of a blank adaptor, which took 30 minutes. The test program consisted of 293 test statements plus loops that brought the total number of tests to 30,000. The total test time was about 7 seconds.

The following examples are brief looks at other circuits of varying complexity.

The board in Figure 3 has 26 inputs and 26 outputs assigned to its logic portion. The logic itself is simply 35 inverters contained within 7 IC's. Programming time: 30 minutes. Device-adaptor preparation time: 30 minutes to wire a blank adaptor. Test statements: 25. Total test time: 10 milliseconds.

The next example, Figure 4, has 11 inputs and 18 outputs, and consists of 27 IC's and 5 discrete components (functionally, decoders, a 3-bit binary counter, a 14-bit flip-flop shift register with parallel output, and 15 read-in gates). Programming time: 8 hours. Device-adaptor preparation time: 30 minutes to wire a blank adaptor. Test statements: 151 (with loops, the total board check consists of 293 tests). Test time: 80 milliseconds.

The final example, Figure 5, is a front-panel assembly consisting of 7 BCD-to-decimal converters, 7 decimal display tubes, six 10-position thumbwheel switches, 24 dpdt pushbutton switches, an 8-position rotary switch, and a dpdt toggle switch. The panel has 32 inputs and 63 outputs. Programming time: 8 hours. Device-adaptor fabrication time: 8 hours to wire a blank adaptor and to construct special cables from panel to adaptor. Test statements: 188. Test time: 3 minutes, including time for the operator to reset controls on the assembly, according to scope-displayed instructions.

Leading parts in the design of the GR 1790 prototype system were played by R. T. Cvitkovitch, M. L. Fichtenbaum, A. W. Winterhalter, and C. Lynn, with R. G. Fulks and D. S. Nixon, Jr. acting in advisory capacities. Development of the version described in the article primarily rested upon the shoulders of Fichtenbaum, P. A. d'Entremont, P. H. Goebel, and J. B. Pennell.

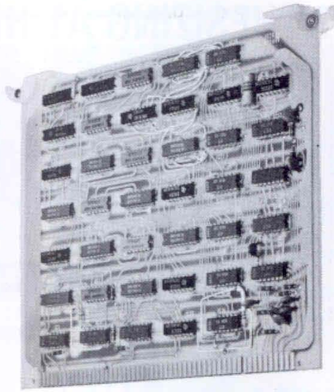


Figure 2. Test time: 7s for 30,000 tests.

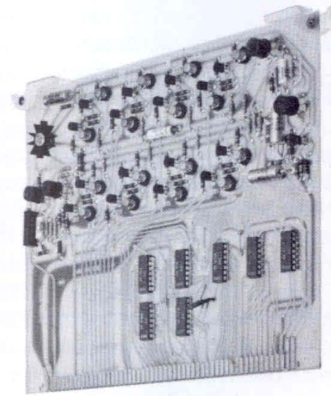


Figure 3. Test time: 0.010s/board.

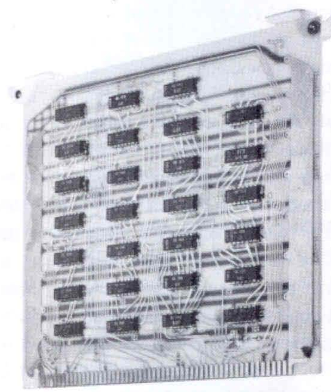


Figure 4. Test time: 0.080s/board.

Complete specifications for the GR 1790 are in GR Catalog U, available shortly, and in a special brochure available on request.

Description	Price in USA
1790 Logic-Circuit Analyzer, console version	**$32,500.00**
Option 1 Rack Version	(no extra charge)
Option 2 Additional Memory	**add 11,500.00**
Option 3 Programmable Logic Levels	**add 9,500.00**
1790-9601 Device Adaptor Kit, without holes for socket, 72 inputs-72 outputs	**110.00**
1790-9602 Device Adaptor Kit, without holes for socket, 96 inputs-144 outputs	**160.00**
1790-9603 Device Adaptor Kit, with holes for socket, 72 inputs-72 outputs	**115.00**
1790-9604 Device Adaptor Kit, with holes for socket, 96 inputs-144 outputs	**165.00**

All prices subject to quantity discount.

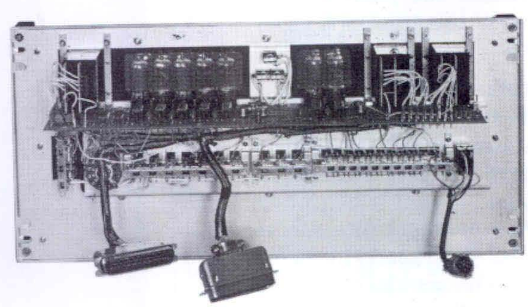
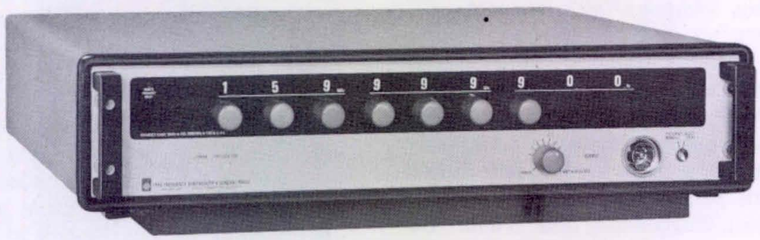


Figure 5. Test time (including control reset): 3 min.

SYNTHESIZING AT HIGHER FREQUENCIES



1165 Frequency Synthesizer

A Bit of Philosophy

The constant pressure of competitive enterprise doesn't encourage complacency in the market place these days. If we mention that the midnight lamps glow frequently at GR, it's not a joke! Our design engineers take their projects in dead seriousness and are under constant pressure to innovate or to improve the existing GR line of instruments. The smiling faces our readers see on the photographs of engineer-authors within the pages of the *Experimenter* usually reflect the pleasant sense of accomplishment when a project is complete. The ultimate test of completion, however, is *acceptance* of the new product by the public, based upon technical performance and *economic cost*.

Cost to the consumer has weighted industrial design considerations very heavily during the past few years and shows promise of even more influence in the decade of the 1970's. With these facts in mind, it is pleasant to announce the availability of another GR instrument, designed for the economy-minded customer. The technical description of the GR 1165 Frequency Synthesizers, which follows, does not emphasize its low cost to any great extent – if our readers are impressed by the specifications and features of this synthesizer, the price information at the end of this article will be a pleasant surprise.

A Bit of Information

The frequency synthesizer is the universal signal source for all applications requiring accurate programmable frequencies. Typical is heterodyning the synthesizer output with another signal carrying intelligence for transmission via radio frequencies or applying the intelligence directly as phase modulation to the synthesizer.[1] The synthesizer can be employed as a heterodyning oscillator for a radio receiver or it can be the driving source for impedance or transmission characteristic measurements in a computer-controlled automatic test system. The synthesizers are ideal for measurement of signal-source stability, and they have found extensive application in nuclear magnetic resonance studies.

The new GR 1165 Frequency Synthesizers extend to 159.9999 MHz the frequency coverage of the GR 1160 series previously announced. The units are remotely programmed by use of BCD coding and are supplied in master and slave versions.

The master unit contains a precision quartz-crystal master oscillator, operating in a temperature-controlled oven, and can be locked to an external frequency standard for greater precision, if desired. A front-panel warning light signals failure to lock to an external source. Provision is made on the master unit for maintaining power to the crystal oven by means of an external battery, in the event of ac power failure.

The slave unit requires an external frequency driving source, provided by the auxiliary 10-MHz output of a master unit or any other precision 5- or 10-MHz source. Each slave unit also has an auxiliary 10-MHz output derived directly from its input; this permits any number of slave units to be cascaded if the first unit in the chain is driven by an external source.

Fundamentals of the derivation of the synthesizer output have been described previously.[2] In the GR 1165 units, one oscillator uses a 3-decade scale-of-"N" phase-lock loop to provide 1000 steps of frequency selection. This technique provides considerable savings in space and production costs as compared with the previously used technique of a single decade of control per locked oscillator.

A Review of Some Characteristics

- Locally or remotely controlled, from 10 kHz to 159.9999 MHz in 100-Hz steps.

[1] "Applications for Coherent Decade Frequency Synthesizers," GR reprint Form No. 3218-A, available upon request to the Editor.

[2] Noyes, A., Jr., "Coherent Decade Frequency Synthesizers," *GR Experimenter*, September 1964.

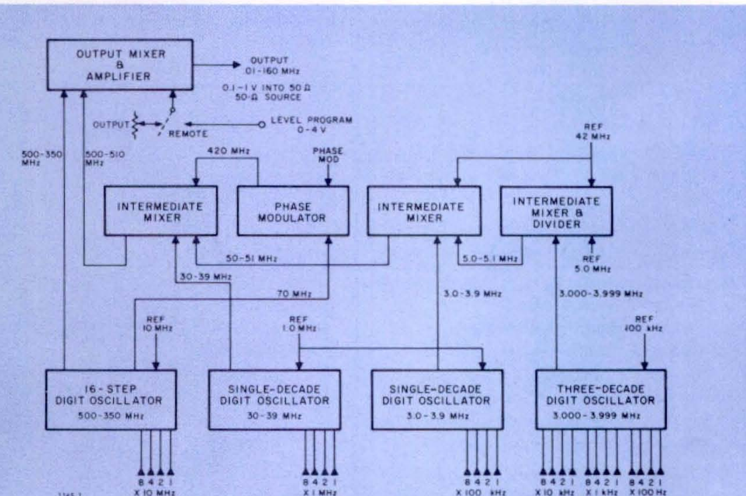


Figure 1. GR 1165 simplified block diagram.

- Auxiliary outputs at 1 MHz and 10 MHz.
- Logic levels are +5 V for logic "0" and +0.5 V or less for logic "1," to facilitate interconnection with other GR instrumentation logic controls.
- All program lines maintain logic "0" unless the external circuit is grounded. A maximum current of 3 mA to ground is required to program a logic "1." These logic levels are compatible with external DTL or TTL IC gates or inverters; a simple external change, using IC inverters, adapts the unit to positive logic-level circuits.
- Externally controlled frequency selection uses 1-2-4-8 binary coded control.
- Can be phase modulated up to ±3 radians at rates from dc to 300 kHz, up to ±1 radian at 1 MHz.
- Continuously energized quartz-crystal-oscillator oven maintains master-oscillator stability of $<2 \times 10^{-10}$ per degree Celsius in 0° to 50° C environment and $\approx 1 \times 10^{-9}$ stability per day after one-month continuous operation.
- Physical height of relay-rack models is only 3½ inches.

– W. F. Byers

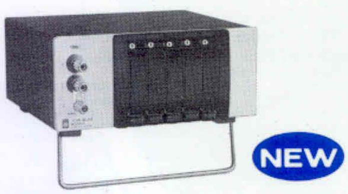
The GR 1165 development was by an engineering team consisting of W. F. Byers, C. C. Evans, G. H. Lohrer, R. L. Moynihan and P. L. Sullivan, assisted by A. E. Carlson and R. J. Hanson, and headed by A. Noyes, Jr., Group Leader.

After graduating from Ohio State University with a BSEE degree, W. F. Byers joined General Radio in 1943 as a development engineer. Presently, he is Group Leader in the GR Signal-Generator Group. He is a Senior Member of IEEE and is registered as a professional engineer in Massachusetts.

Complete specifications for the GR 1165 appear in GR Catalog U, available shortly.

Catalog Number	Description	Price in USA
	1165 Frequency Synthesizer, master version	
1165-9720	Bench	**$5900.00**
1165-9721	Rack	**5900.00**
	1165 Frequency Synthesizer, slave version	
1165-9722	Bench	**5300.00**
1165-9723	Rack	**5300.00**

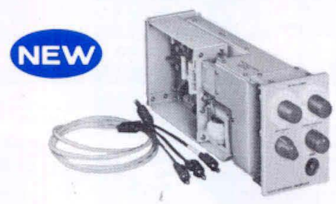
All prices subject to quantity discount.



The trend toward miniaturization of test equipment is exemplified by the new GR 1436 Decade Resistors, available in two values: 111,110 Ω and 1,111,100 Ω, with smallest steps of 1 Ω and 10 Ω respectively. In addition, control by convenient lever switches facilitates rapid adjustments. Contacts are made of solid silver-alloy; the units of higher resistance value are wound with Evanohm* wire, the units of lower resistance value with Manganin** wire. Both models of the GR 1436 are available without cabinets for custom installations; inquiries are invited. Physical size of the new units is 8-1/2 × 3-7/8 × 8-5/16 in. (220 × 99 × 213 mm). The new resistors were developed by W. J. Bastanier, development engineer in the GR Component and Network Testing Group.

*Registered trademark of Wilbur B. Driver Co.
**Registered trademark of Driver-Harris Co.

The introduction of new coaxial capacitance standards GR 1405-A (20 pF) and GR 1405-B (10 pF) extends the existing GR line of coaxial capacitance standards. Now available are units ranging from 1 pF to 20 pF, terminated in the GR900® precision connector. Development of the new capacitance standards was the responsibility of J. Zorzy, Group Leader of the GR Microwave Group.



The GR 1522-P2 Differential Preamplifier provides for operation of the GR 1522 Recorder from ungrounded signal sources. Its differential input will handle a wide range of voltage and current measurements. Common-mode rejection up to 180 dB is a feature at inputs up to 500 volts. The preamplifier was developed by J. M. Steele, development engineer in the GR Acoustics/Signal Analysis Group.

Complete specifications for the above units appear in Catalog U, available shortly.

Catalog Number	Description	Price in USA
	1436-M Decade Resistor, 111,110 Ω	
1436-9700	Bench Model	**$210.00**
1436-9701	Rack Model	**245.00**
	1436-P Decade Resistor, 1,111,100 Ω	
1436-9702	Bench Model	**230.00**
1436-9703	Rack Model	**265.00**
1405-9704	**1405-A Coaxial Capacitance Standard,** 20 pF	**85.00**
1405-9703	**1405-B Coaxial Capacitance Standard,** 10 pF	**85.00**
1522-9602	**1522-P2 Differential Preamplifier**	**475.00**

All prices subject to quantity discount.

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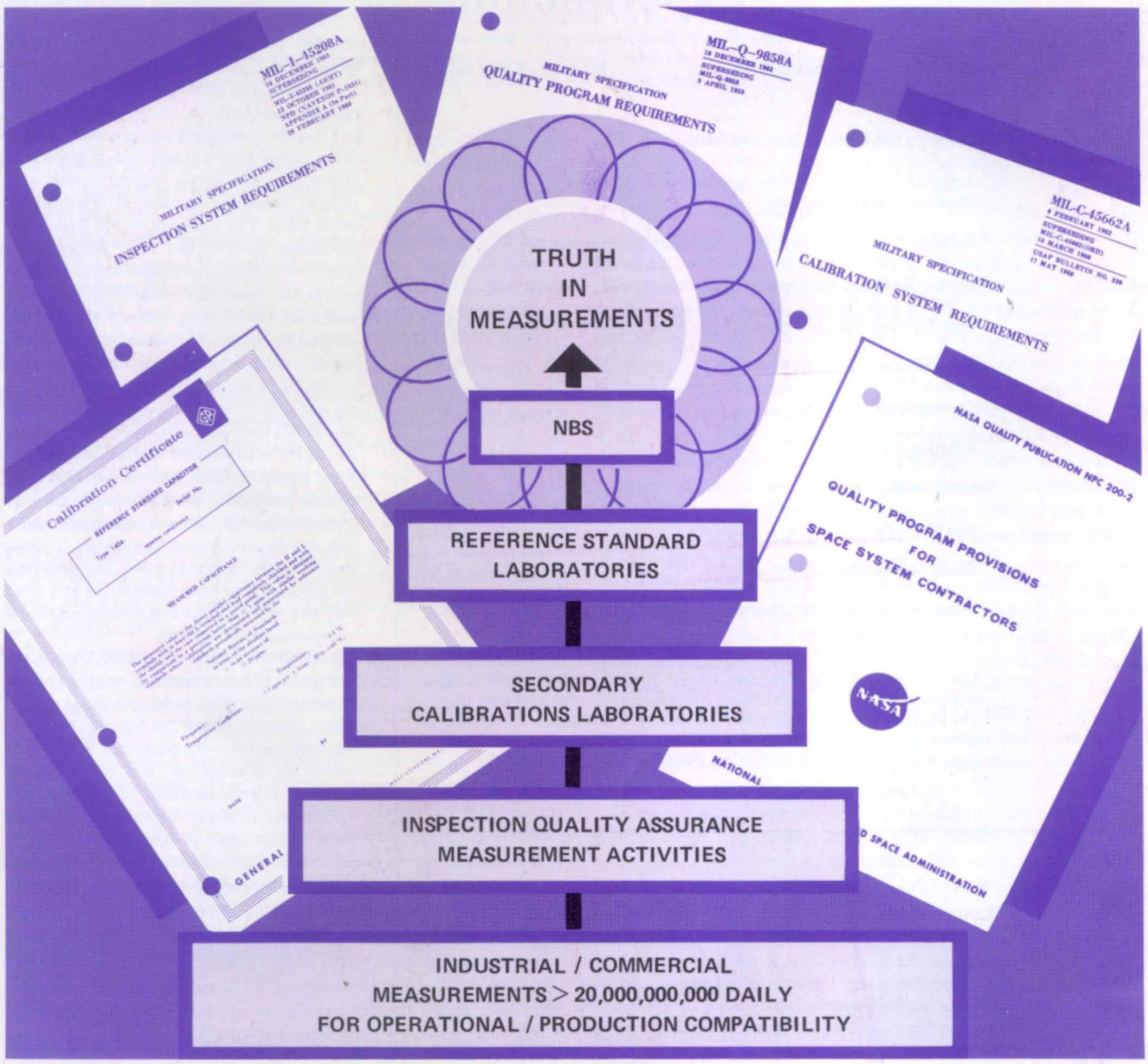
NUMBERS 3, 4, 5, 6
MARCH – JUNE 1970



THE
GENERAL RADIO
Experimenter

VOLUME 44 NUMBERS 3-6 MARCH/JUNE 1970

The *General Radio Experimenter* is mailed without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., West Concord, Mass. 01781.

THE COVER Our government employs a system of measurement-accuracy control expressed in contractual specifications as "traceability to the National Bureau of Standards." It is a philosophy based upon one principle: interchangeability and compatability of industrial products are attainable through accurate measurements based upon physical standards common to all users. A discussion of the subject appears on page 6.

Our last issue contained a lead article devoted to the economic justification of a large capital investment in automated test equipment. The purposes in writing the article were twofold: 1) to ally ourselves with our readers in the common problem of justifying capital expenditures, and 2) to demonstrate by example an economic reasoning process proven acceptable to one management team. The response to the article was somewhat startling but most pleasing. Requests for a reprint of a talk referred to in the article depleted our supply within three weeks of the mailing date of the *Experimenter*. It has been necessary to make more copies available for the continuing demand.

All of which proves that most engineers are ***not*** unmindful of the impact upon their company's financial position of their requests for capital expenditures for instrumentation. They recognize that their expenditures are directly related to the profitability of their company and are willing to act as financially responsible, reasonable individuals — when given direction and guidance in procurement. We are happy to have contributed in a small way to increasing the engineering value and knowledge of our readers.

The next logical step is to discuss the economics of instrument utilization. We hope to present, shortly, some thoughts, experiences, and advice on other forms of engineering management — the control of instruments within an organization, and the economic waste of mismanagement. The subject is a vexing one in many companies, large and small. It has been of great concern to our government, which has a tremendous investment in test instrumentation in its own and in its contractors' plants.

Perhaps you have some thoughts on the subject? If so, we'll be glad to hear from you.

C E White

C. E. White
Editor

Toward a More Useful Reference Standard Resistor

The Electrostatic System of Units and Measurements provides a method of establishing the value of a standard resistor directly from the value of the computable capacitor developed by Thompson-Lampard in Australia. The value for a resistor which more closely approaches the capacitor's impedance reduces the multiplication factor for calibration purposes; the comparison-measurement accuracy is thereby increased up to some definite value for the resistor. For practical reasons, the compromise value of 10,000 ohms has been established recently for a new standard of resistance. Development work by General Radio has produced a resistor with the high stability of the Thomas 1-ohm resistor and with a better temperature coefficient. Our new and different design approach is described in this article.

A number of articles have been written on the advantages of a 10,000-ohm reference standard resistor in comparison to a 1-ohm standard. Some writers have gone so far as to assume that a 10,000-ohm resistor will replace the famous Thomas-type 1-ohm standard. I'm afraid we will have to wait a long time – perhaps 30 years – before this can be proven. But in the meantime, let's see what some of the advantages are to calibration laboratories presently dependent upon the 1-ohm reference standard.

- The 10,000-ohm standard reduces the number of transfer steps required to measure resistors in the range from 1,000 ohms to the highest calibration range required.
- The 10,000-ohm value is much closer to, and can be more easily derived from, the very stable value of the Thompson-Lampard computable capacitor.
- The 10,000-ohm value reduces to a minimum the errors due to lead and contact resistance and thermal voltages, yet the value is not so high that shunt leakage resistance is a major problem.
- The 10,000-ohm value presents a better impedance match to modern null detectors, giving better signal-to-noise ratio, higher detector sensitivity, and reduced power dissipation in the bridge circuit.

What makes the Thomas standard resistor so stable and therefore universally accepted? It consists of a coil of bare, heavy Manganin* wire that has been heat-treated at 500°C approximately. Next, this coil is placed very carefully on an enameled brass cylinder (any mechanical stress or strain is avoided), then it is hermetically sealed in dry air or nitrogen. This is, of course, a very simple description of the production process; the main point of interest is that the wire is heat-treated at a high temperature *after* it has been formed into a coil. The coil is supported by a heat-conducting form with a temperature coefficient practically the same as the Manganin. There is a disadvantage in this standard, however. Its high temperature coefficient requires extreme control of the operating environment.

Contrast the production process above with the usual construction of a 1,000-ohm or 10,000-ohm resistor. Here a much thinner, insulated wire is used which cannot be wound into a self-supporting coil. It is necessary to wind the coil upon some form of substrate, and to wind it loosely to avoid excessive stress or strain in the wire. Some manufacturers use bobbins for support, others use cards; every unit is heat treated cyclically in an attempt to relieve any built-in stresses. The range of temperature for treatment may go as high as +150°C and as low as −80°C. This process results in a good resistor but not equal to the Thomas-type, since the temperature range used is not sufficient to relieve all the stresses and strains in the wire. Eventually the stresses retained result in a positive resistance drift with time if, for example, Evanohm** wire is used.

Another factor that affects resistance stability to a certain extent is the coating on thin wire. As the coating dries and hardens it produces a restricting or "choking" effect on the wire, and its resistance increases. When the wire is placed in an oil bath, the coating softens and the wire resistance decreases. Obviously, it is desirable to eliminate the coating on wires if a stable standard is under consideration. This was done in the design of the 10,000-ohm GR 1444 Reference Resistance Standards.

Basic construction of the GR 1444 uses an insulated metal substrate 3½ × 3 inches, having a coefficient of thermal expansion similar to that of Evanohm. The resistors are wound with bare Evanohm wire to a nominal value of 5000 Ω, and heat-treated in an inert atmosphere at approximately 500°C. We find that this treatment gives complete relief from stresses and strains in the wire. There is an additional benefit: the resistor's temperature coefficient can be adjusted to a very close tolerance *after* the resistor is wound. This is possible because the electrical properties of Evanohm change at ele-

*Registered trademark of Driver-Harris Company.

**Registered trademark of Wilbur B. Driver Co.

1444-B 10-kΩ standard, for use with oil bath, shown removed from carrying case.

1444-A 10-kΩ standard with temperature sensor, shown with carrying case.

vated temperatures as a function of time (refer to note at end of article).

Upon completion of the heat treatment, two of the resistors are mounted upon a heavy brass plate on either side of a center shield, to which is soldered a thermometer well. The GR 1444-A standard includes a 10,000-ohm temperature-sensor resistor network, wound upon the center shield. The network has a temperature coefficient of 1000 ppm/°C and is in close thermal contact with the standard resistor elements. The GR 1444-B standard does not contain a temperature sensor since it is intended for use in an oil bath. An individual temperature-correction chart is supplied for either type standard.

The complete GR 1444 unit is welded into a 1/16-in. stainless steel container, which is evacuated, filled with dry nitrogen, and then sealed. We chose not to use an oil-filled enclosure because no oil is completely free of water. Any water content would cause oxidation of the wire. The effect of oxidation increases with decreased wire diameter. The GR 1444 uses wire of 0.002-in. diameter, and an oxide film of only 1 Å (0.00394 μin.) will produce a resistance change of 8 ppm. An extended laboratory test of a GR 1444-A standard demonstrated the stability of the unit over a two-year period (Figure 1).

Since a useful standard of any type ideally should be comparatively unaffected by a hostile environment, we

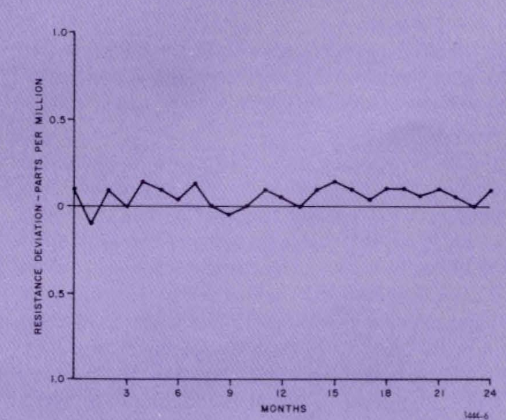


Figure 1. Typical long-term stability of GR 1444 resistors.

investigated the effect of shock upon the GR 1444-A. The unit was dropped three times, from a height of 36 inches, upon a concrete floor. The result of several tests was a resistance change of −0.1 to −0.2 ppm. Obviously, we don't imagine that you would, in normal practice, deliberately mistreat a standard in this manner. If, however, it were to happen, the value of the standard would not be impaired. By the way, after a two-day rest period the measured change was found to be less than ±0.1 ppm. With proper design of transport packaging, the unit will withstand shipment for calibration or any other purpose.

Of interest to specification-minded professional people are two other tests we conducted. The ambient pressure was varied from 1000 microns to 3 atmospheres (10^{-3} to 2280 mm Hg), and the resistance change was less than 0.1 ppm. Leakage resistance between the resistor and case was measured at 4×10^{12} Ω at 35% RH and 2×10^{12} Ω at 94% RH.

—W. J. Bastanier

The author wishes to acknowledge with deepest gratitude the cooperation of Dr. C. D. Starr and of the Wilbur B. Driver Company in permitting use of the heat-treatment techniques developed by them.

Catalog Number	Description	Price in USA
1444-9700	**1444-A Reference Resistance Standard,** 10 kΩ, with sensor	**$600.00**
1444-9701	**1444-B Reference Resistance Standard,** 10 kΩ, without sensor	**600.00**

Prices subject to quantity discount.

Some Technical Notes

Starr describes the useful functional changes of Evanohm wire with time as it is exposed to high temperatures.[1] Completely annealed Evanohm wire is commercially available with a temperature coefficient of +50 ppm/°C. During the heat treatment process the temperature coefficient decreases, passes through zero, reaches a maximum negative value of about −35 ppm/°C, and then passes through zero a second time to become positive once again (Figure 2). The first zero-crossing point is considered to give the most stable condition of the wire.

At the same time that the temperature-coefficient effects are observed, the resistivity of the wire also is changing. Starting at about 730 ohms/cir mil ft, the resistivity rises to approximately 800 ohms/cir mil ft when the temperature coefficient makes the first zero crossing. Continued heat treatment causes the resistivity to reach a maximum value and finally it decreases slightly.

The heat treatment of GR 1444 standards is such as to produce a temperature coefficient of ± 0.1 ppm and a resistivity of approximately 800 ohms/cir mil ft.

[1] "Properties of Wire for Resistors," C. D. Starr, *Materials Research & Standards*, September 1966.

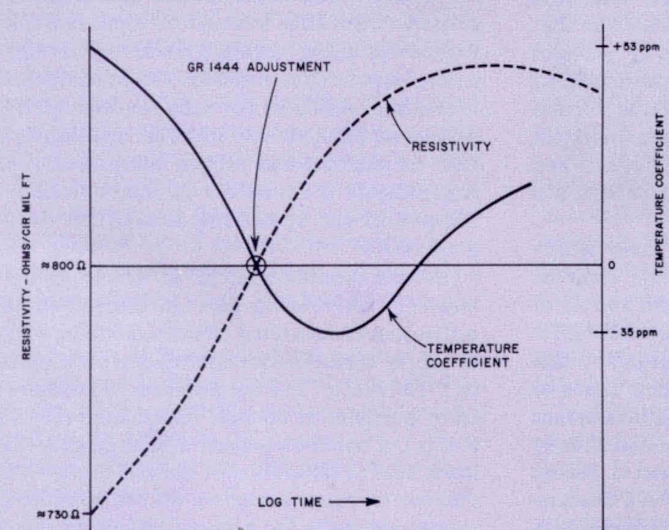


Figure 2. Functional changes of Evanohm wire.

Accuracy Traceability

Its Impact upon Instrument Manufacturer and Customer

Dedicated people, experience, and technical knowledge assure operational integrity of measurement standards and calibration activities. These ingredients, however, are not sufficient to assure measurement compatibility with similar activities, nor with the National Bureau of Standards. Required is some sort of measurement interface between production measurements and national standards. The Department of Defense calls this interface "traceability" but does not define it sufficiently or comprehensively. This article relates some background of the traceability problem and our analysis and evaluation of traceability philosophy, and it introduces a fourth path to establishment of traceability.

Shortly after issuance (9 April 1959) by the U.S. Department of Defense of military specification MIL-Q-9858, "Quality Control Requirements," a chorus of voices raising questions was heard. Particularly of interest to all workers in the field of measurement standards and calibration was the clearly expressed requirement stated under the topic heading "Traceability of Calibration:" "In the Zone of Interior, Hawaii and Alaska, measuring and test equipment shall be calibrated with measurement standards, the calibration of which is traceable to the National Bureau of Standards . . ."

Immediately apparent to many metrologists was the impracticality of attempting to establish a traceable pattern of accuracy of state-of-the-art measurements for which the Bureau had no physical standards. It was apparent also that many measurements were capable of evaluation by the ratio type of self-calibration techniques. Finally, measurements based upon independent, reproducible standards derived from accepted values of natural physical constants could gain no further accuracy, if performed originally by qualified personnel, by comparison with similar standards at NBS, and in fact would only increase operational costs.

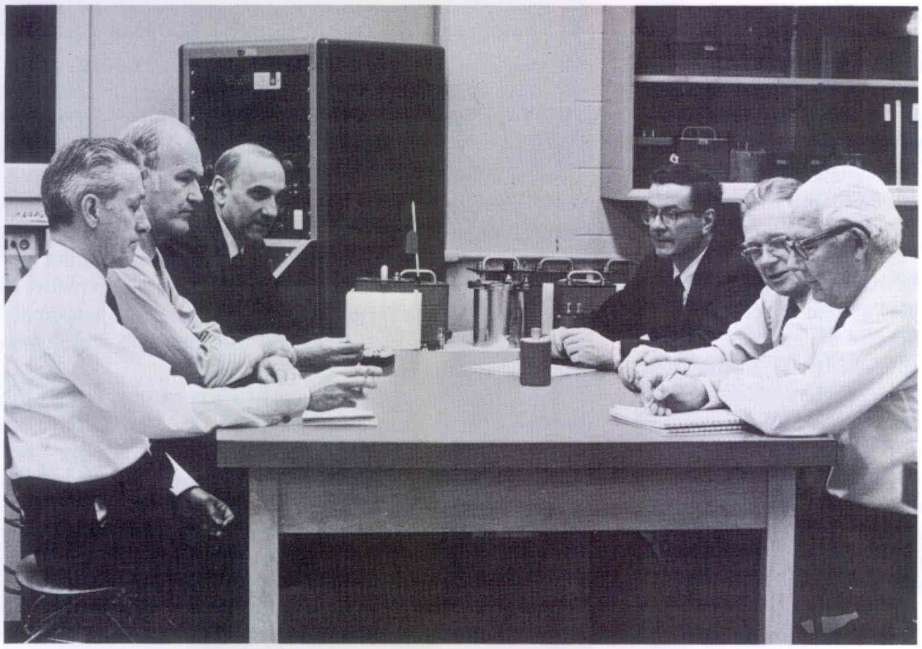
Official voices of protest were raised within many organizations, e.g., Aerospace Industries Association, National Conference of Standards Laboratories, Instrument Society of America, and Scientific Apparatus Makers Association. Instrument manufacturers found themselves besieged by customers requesting guidance or advice concerning means or techniques that would be acceptable to quality-assurance inspectors in cases involving standards not calibrated directly by NBS. Among the many voices was that of Ivan G. Easton of General Radio, who expressed these feelings[1] at a measurements symposium sponsored by American-Bosch Arma Division:

"The present hysteria about traceability to NBS is disturbing. The word hysteria is perhaps a bit strong but we have seen situations develop in the past year in which the use of the word does *not* constitute excessive hyperbole. While the intent of the traceability program is crystal clear, the operation of the program needs to be placed in proper perspective. The concept that all calibrations be based ultimately on NBS certification is a commendable one, but the means of achieving this result requires scrutiny. The interpretation of the traceability requirement, in some circles, has led to the demand that a manufacturer state the date of the last NBS certification of the standard against which the item being sold was checked. This request, while at first glance innocent and entirely reasonable, is based on a misconception of a standards structure. A standards laboratory consists of *more* than a set of standards certified by NBS on some particular date. A calibration depends not only on standards, but also on people, on methods and, above all, on integrity . . . . I believe that clarification and proper interpretation of traceability requirements is a problem for the immediate future at the commercial and operational level of our national standards program."

Several months later in an article in the *GR Experimenter*,[2] Mr. Easton gave an illustration of one of the difficulties encountered by manufacturers who are requested to supply corroborative information pertaining to "traceability." He wrote, "In a few instances . . . we have been requested to supply more detailed information (pertaining to traceability). A common request in such cases is that we state the latest date of calibration of our standard by NBS. There are a number of reasons why we do not consider this to be the proper approach for a manufacturer of standards, and we have consistently avoided the supplying of such information.

[1] Easton, I. G., "The Measurement and Standards Problems of Tomorrow," January 25, 1960, unpublished.

[2] Easton, I. G., "Standards and Accuracy," *GR Experimenter,* June 1960.



Editor White (left) discusses problems of traceability with GR engineers (from left to right) J. Hersh, J. Zorzy, H. Hall, R. Orr, and W. Bastanier who developed the GR 1444 standard resistor described on page 3.

The most significant objection to this procedure is that it constitutes an oversimplification and does not recognize the true nature of a standards structure . . . we believe quality assurance cannot be obtained from any single detail such as the date of an NBS calibration . . . "*

Probably the most important point in the consideration of traceability as a required component in a measurement system is the fact that traceability itself does not assure accuracy in measurements. Errors exist within a measurement system because of many reasons, including poor measurement techniques, carelessness or inexperience of measurement personnel, physical changes and consequent drift of instrumentation operating parameters.

We can also consider *degree* of traceability. There is no way in which traceability can be *quantified*. We agree with Lord Kelvin, if you can't put a number on a quantity it can't be measured. True traceability, instead, is an expression of the operating philosophy and integrity of a measurement activity. Its introduction into a measurement system is an attempt to promote honesty in measurements and to supply a base point (an NBS calibration) from which all like measurements can be derived in confidence.

The National Bureau of Standards, focal point of attention by government and industry alike, felt compelled to express its position in the traceability debate and did so. In a presentation at Boulder, Colorado, 23 January 1962,[3] W. A. Wildhack, Associate Director of the NBS Institute for Basic Standards offered these words:

*Editor's Note: This information *is* available, however, upon written request to GR's Standards Laboratory.

[3]Wildhack, W. A., "Statement on Traceability," Report of Workshop on Measurement Agreement – National Conference of Standards Laboratories, Boulder, Colorado, 23 January 1962.

"The National Bureau of Standards has received numerous inquiries concerning the meaning of the term 'traceability' as used by Department of Defense agencies in contractual documents and elsewhere, in stating requirements pertaining to physical measurements.

"Without further definition, the meaning of this term is necessarily indefinite as applied to relationships between calibrations by NBS and measurement activities of manufacturers and suppliers. 'Traceability' is not given any special meaning by the National Bureau of Standards, and information as to possible special meanings of the term in military procurement activities cannot, of course, be supplied by NBS. Where questions arise, it is suggested that they be directed to the cognizant military inspection or contracting agency."

Obviously, Mr. Wildhack had made the Bureau's position clear in the matter of traceability, but the position of the manufacturer and user of instruments remained as precarious as before. Fortunately, the specification that had caused the trouble was improved by a supplement specification, MIL-C-45662-A, issued 9 February 1962. Bowing to the logical and technically correct points raised by metrologists, the government agreed that the traceability provision could be established by any of three methods:

a. Intercomparison directly with, or through, an echelon of standards laboratories to the National Bureau of Standards,

b. Intercomparison with an independently reproducible standard acceptable to the Bureau,

c. Derivation of accuracy by use of a ratio type of self-calibration technique acceptable to the Bureau.

The last mentioned method is undoubtedly a most useful tool for any calibration activity. It is recommended by the

Bureau because of its cost-saving features and the reduction of the heavy workload at the Bureau by any work performed by laboratories themselves. Unfortunately, there are many aspects to this method, not known to all measurements workers. Questions raised by General Radio customers continually remind us of this fact. We have been encouraged to present to our readers NBS-acceptable techniques practiced by our engineering staff, which are available to many measurement laboratories.

We approached GR's Henry P. Hall for advice and assurance concerning impedance-calibration techniques that are technically sound and display more than a modicum of common sense. His reaction was immediate and positive. He pointed out the fact that the Bureau itself, possessing a minimum of standards, also faces the problem of traceability on an international scale. Through the years the Bureau has established techniques similar to those listed above as a, b, and c. For more efficient use of its limited budget, facilities, and manpower, it has been forced to develop dependable calibration techniques using other than comparison methods. Most important, techniques so developed are available for use by any technically qualified calibration activity.

From a series of round-table discussions with members of GR's Component and Network Testing Group came information on one of these techniques, technically sound and requiring no recurrent calibration services from the Bureau. Briefly stated, the technique merely involves the application of known frequency correction factors to standards. We think of the technique as Self-Calibration by Frequency Translation. It has been applied vigorously at GR (and NBS) for a number of years. Its use is particularly emphasized in calibration of impedances for one very important reason:

"It has long been known that capacitors are superior to either inductors or resistors for use as impedance standards at high frequencies, because changes in their effective values with increasing frequency can be more accurately and more easily evaluated than such changes in either resistors or inductors."[4]

Some administrative problems of the calibration laboratories emanate from the assumption that all measurements must be *directly* traceable to the Bureau to assure measurement accuracy. Overlooked by many laboratories is the fact that they possess measurement capabilities of such a nature as to provide self-help far greater than they realize. They simply apply known data given them by the standards and instrument manufacturers!

Take, for example, the request by a customer to solve the problem of calibrating a 1-pF capacitor at 1 MHz without the necessity of going directly to the Bureau. He already has a certificate of calibration at 1 kHz, routinely supplied at the time he made the original purchase. Since then, the measurement range in his company's work has expanded into the high-frequency field. His operating budget is nominal and he is interested in keeping his overhead down. The answer he receives from GR is covered in the following article by Mr. Hall.

[4] Huntley, L. E., "A Self-Calibrating Instrument for Measuring Conductance at Radio Frequencies," *NBS Journal of Research-C*, April-June 1965.

The Editor acknowledges with gratitude the work of Mr. Hall in preparing the above article and is equally appreciative of the constructive discussions with J. F. Hersh, R. W. Orr, and J. Zorzy.

Another Traceability Path for Capacitance Measurements

by Henry P. Hall

The National Bureau of Standards recently announced new services for the rf calibration of impedance standards fitted with precision 14-mm coaxial connectors, like several manufactured by GR.[1,2,3] High frequency bridges and twin-T circuits developed at the Bureau are used for the measurement of resistance, capacitance, and inductance at listed frequencies of 0.1, 1.0, and 10 MHz.[4,5] Anyone who wants direct NBS "traceability" of these standards may send them to NBS for calibration at these test frequencies. This is a significant advancement in the measurement art.

The new service, however, raises an important question: "When are these *direct NBS calibrations* necessary?" This can be considered from two points of view: "When are these calibrations required to obtain a specific accuracy with high confidence?" and "When are they required to prove traceability as defined (somewhat loosely) by the DOD?" The two points of view diverge by varying degrees, unfortunately, particularly when we consider to whom we are trying to prove traceability. We think that the question can be answered satisfactorily for rf-capacitance calibrations, if you are interested in obtaining a *given required accuracy* through use of frequency-correction data. We hope that reason will prevail among metrologists and that the same criteria used in deciding what is sufficiently accurate can also be used to show traceability, for the benefit of quality-assurance inspectors.

The frequency corrections described below were used with apparent acceptance[6] before the new services were available. They still must be used at frequencies between those for which calibrations are now routinely provided.

Actually, all the new NBS rf-impedance calibrations, *R*, *L*, and *C*, are based on the frequency characteristics of air capacitors. While air capacitors vary with frequency in a very simple manner (see below), resistors and inductors vary with frequency in too complex a manner to be as predictably accurate in the rf range. Therefore, these new NBS bridges and twin-T's relate *L* and *R* back to fixed capacitors and capacitance differences in variable air capacitors.[7,8]

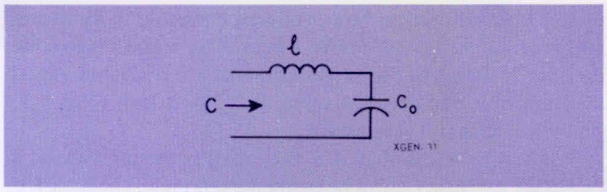


Figure 1. Simple equivalent circuit for two-terminal air capacitor.

RF Characteristics of Air Capacitors

The simple equivalent circuit of Figure 1 is surprisingly good for representing the effective capacitance of a two-terminal air capacitor. The effective capacitance is

$$C = \frac{C_o}{1 - \omega^2 l C_o} \qquad (1)$$

where C_o is the low-frequency value of capacitance. This equivalent circuit does not include series resistance, which does not appreciably affect the capacitance (even the parallel capacitance) if it is reasonably small. Neither does it include the effects of distributed inductance and capacitance. Also, at high frequencies the skin depth causes a slight reduction in inductance. These effects can be considered as variations in the value of l, with C_o being considered a constant (as long as all measurements are made at low humidity, $< 40\%$ RH).

The simplest method of establishing l is to use a grid-dip meter to determine the resonant frequency of the shorted capacitor. This method, together with the above formula, was recommended for many years by NBS[6] to make rf-capacitor calibrations. For capacitors fitted with binding posts or banana-pin terminals, this is very easy to do, although we have to decide how much of the total inductance is in the shorting connection. When coaxial connectors are used, we remove the case to allow coupling to the meter coil.

The value of l obtained from a resonance measurement differs slightly from a lower-frequency value mainly because of the distributed parameters. The worst case would be that of a uniform line whose effective capacitance could be written:

$$C = \frac{C_o}{\left[1 - \omega^2 l C_o - \dfrac{\omega^4 l^2 C_o^{\,2}}{5} - \dfrac{2\omega^6 l^3 C_o^{\,3}}{35} \ldots\right]} \qquad (2)$$

The derivation is a power series in increasing powers of $\omega l C_o$. The first term is the important one for corrections at frequencies well below resonance. When $\omega^2 l C_o = 10\%$ the next term is 0.2%, and each succeeding term is much smaller still. This effective l is 1/3 the value of the total inductance. If C_o is assumed constant, the effective inductance determined by a resonance measurement would be higher than l by the factor $\frac{12}{\pi^2} = 1.21$ or 21%.

In large air capacitors (50 pF to 1000 pF) with parallel-plate construction,[1] most of the inductance is in the coaxial connector and its connection to the "stack" of plates. Most of the capacitance is in the stack itself so that the lumped equivalent circuit is a good approximation. The slight difference between the resonant value of l and the desired correction value is small, less than the uncertainty of the measurement. Many measurements were made at General Radio to establish this difference so that a more accurate value of l could be determined.[9] The values obtained agreed closely with NBS determinations (within 10%).

The smaller capacitors (1 to 20 pF) have an appreciable distributed capacitance; consequently, the value of resonant inductance will be in error, but by less than 21%. The effective value for these units was determined by a slotted-line technique.[9] These values, supplied by General Radio for our line of coaxial standard capacitors, are not for use above a specified frequency (250 MHz for the 20- and 10-pF units and 500 MHz for the smaller units). This restriction is mainly because of the variations due to distributed parameters. (Correction curves are supplied with units at higher frequencies.)

An important point, which is verified by the many measurements described above, is that different capacitors of the same value and construction have equal inductance, well within the accuracy of the determinations. This is not surprising. The inductance is mostly in the coaxial line of the

H. P. Hall is a graduate of Williams College (AB – 1952) and M.I.T. (SB and SM in Electrical Engineering – 1952). His first association with General Radio was as a co-op student in 1949, and he assumed full-time duties as development engineer in 1952. He became Group Leader of the Impedance Group in 1964 and presently is an Engineering Staff Consultant. He is a member of Phi Beta Kappa, Eta Kappa Nu, Tau Beta Pi, and Sigma Xi. He is a Senior Member of the IEEE and a member of PMA and has served on several committees.

precision connector, or in the stamped metal pieces that make connections to the plates. Measured variations in l, between similar units, are generally less than 3%, and allowing for a 10% change would be very conservative.

When are Calibrations Necessary?

Equation (1) can be rewritten:

$$C = \frac{C_o}{1 - \omega^2 (l_s \pm \Delta l) C_o} \tag{3}$$

where l_s is the specified value of inductance and $\pm\Delta l$ the possible deviation in this value. This deviation is due to error in determination of the inductance and its changes with frequency or between units. The low-frequency value C_o can be easily determined to 0.01% ± 0.005 pF on a GR 1615-A Capacitance Bridge by use of the coaxial adaptor.[9] This value can be used as long as $\omega^2 l_s C_o \times 100\%$ is less than the accuracy required. Corrections should be used above this frequency, and may be used with confidence as long as $\omega^2 \Delta l C_o \times 100\%$ is less than the accuracy required. If greater accuracy is required an NBS calibration should be made. Remember, however, that NBS accuracy is based on similar measurement techniques, subject to similar uncertainties.

The question now is: what is the value for Δl? A value of $\Delta l = 0.2\ l_s$ can be used with confidence as long as the total correction is 10% or less, the maximum frequency for the low values of capacitance is not exceeded, and humidity is reasonably low. Typically, errors will be substantially less. We suggest that calibration be considered traceable to an uncertainty of $\omega^2 N l_s C_o \times 100\%$ where N is some reasonable fraction. If we agree that measurements are traceable, based on some very conservative value of N, even as high as 0.5, this approach is still very useful.

We wish to stress that it is the small *correction* which is important because C_o can be easily determined. The high-frequency capacitance value will vary with time, temperature, or shock because C_o changes with these effects. But unless there is catastrophic damage, the inductance will not change enough to be perceptible. Therefore, *differences* between high-frequency and low-frequency calibrations should be recorded and once made on a given unit need not be repeated. This point should be appreciated also by quality-assurance inspectors.

Conclusion

The principle of applying conservative tolerances to frequency corrections has relevance to frequency translation of other types of impedances. Capacitors with dielectric materials other than air have additional sources of deviations, but these can be determined within definite limits and included in the over-all uncertainty. Frequency corrections are significant for larger capacitors at much lower frequencies, and the corrections greatly extend the useful frequency range of the capacitors. This principle may be applied also to inductors and resistors, even though their equivalent circuits are more complex.

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[1] Orr, R. W., "Capacitance Standards with Precision Connectors," *GR Experimenter*, September 1967.

[2] Orr, R. W., and Zorzy, J., "More Coaxial Capacitance Standards," *GR Experimenter*, May 1968.

[3] Orr, R. W., "Stable Series of Coaxial Resistance Standards," *GR Experimenter*, March/April 1969.

[4] Jones, R. N., and Huntley, L. E., "A Precision High-Frequency Calibration Facility for Coaxial Capacitance Standards," *NBS Technical Note 386*, March 1970.

[5] Mason, H. L., "Calibration and Test Services of the NBS," *NBS Special Publication 250*, 1968 Edition.

[6] Jones, R. N., "A Technique for Extrapolating the 1-kc Values of Secondary Capacitance Standards to Higher Frequencies," *NBS Technical Note 201*, November 5, 1963.

[7] Huntley, L. E., "A Self-Calibrating Instrument for Measuring Conductance at Radio Frequencies," *NBS Journal of Research-C*, April-June 1965.

[8] Field, R. F., and Sinclair, D. B., "A Method for Determining the Residual Inductance and Resistance of a Variable Air Condenser at Radio Frequencies," *Proceedings of the Institute of Radio Engineers*, February 1936.

[9] Zorzy, J., and McKee, M. J., "Precision Capacitance Measurements with a Slotted Line," *GR Experimenter*, September 1967.

For more information related to the background of this article, the reader is referred to:

Woods, D., "A Precision Dual Bridge for the Standardization of Admittance at Very-High Frequencies," *IEE Monograph 244R*, June 1957.

Woods, D., "Admittance Standardization and Measurement in Relation to Coaxial Systems," *IRE Transactions on Instrumentation*, September 1960.

Huntley, L. E., and Jones, R. N., "Lumped Parameter Impedance Measurements," *Proceedings of the IEEE*, June 1967.

Recent Technical Articles by GR Personnel

"Computer-Controlled Testing Can Be Fast and Reliable and Economical Without Extensive Operator Training," M. L. Fichtenbaum, *Electronics*, 19 January 1970.*

"Using Stroboscopy," C. E. Miller, *Machine Design*, 30 April and 14 May, 1970 (two parts).*

"How Do You Buy a Counter?," R. G. Rogers, *Electronic Products*, 15 March 1970.

"New Fused-Silica-Dielectric 10- and 100-pF Capacitors and a System For Their Measurement," D. Abenaim and J. F. Hersh.**

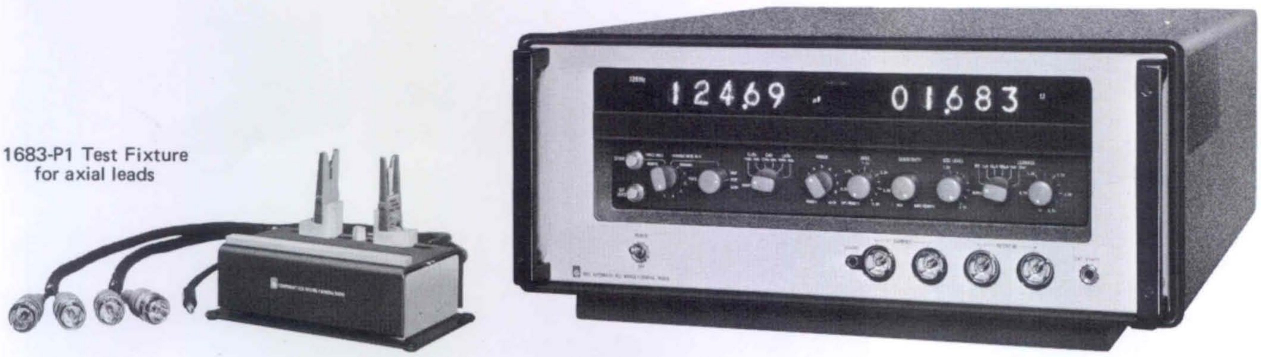
"A Four-Terminal, Equal-Power Transformer-Ratio-Arm Bridge," H. P. Hall.**

"Standardization of a Farad," H. P. Hall.***

*Reprints available from General Radio.

**Presented at the Conference on Precision Electromagnetic Measurements, Boulder, Colorado, June 2-5, 1970.

***Presented at the Annual Conference of the Precision Measurements Association, Washington, D. C., June 17-19, 1970.



1683-P1 Test Fixture
for axial leads

1683 Automatic RLC Bridge

FIVE-TERMINAL AUTOMATIC RLC BRIDGE

To say that General Radio customers aided in the design of the GR 1683 is an understatement. Virtually all the bridge features are the result of feedback from customers of the earlier GR 1680.[1]

WHAT YOU ASKED FOR

First and foremost, many of you wanted to measure higher-valued electrolytic-class capacitors, which calls for better bias capabilities, short-circuit protection, and leakage-current measurements. Some needed to express loss as equivalent series resistance. Still others needed to measure inductors or resistors. All, of course, needed faster measurement rates.

[1] Fulks, R. G., "Automatic Capacitance Bridge", *GR Experimenter,* April 1965.

WHAT YOU'RE GETTING

What you asked for – *plus fast balancing*. Faster measurement rates required a new balancing technique. Many schemes were designed, simulated by a computer model, and analyzed. The rejected methods were either too slow, too complicated, or too expensive. The selected technique (Figure 1) provides balance time of approximately 100 periods of the bridge test frequency.

Fast Balancing

We use three techniques - all time savers. The first is an error-counts-controlled clock rate by which the bridge converges at variable rates – quickly for a large unbalance and more slowly for a small unbalance.

The second technique compensates for the time delay in the filter through use of an anticipator circuit. Its operation is illustrated in Figure 2. Assume

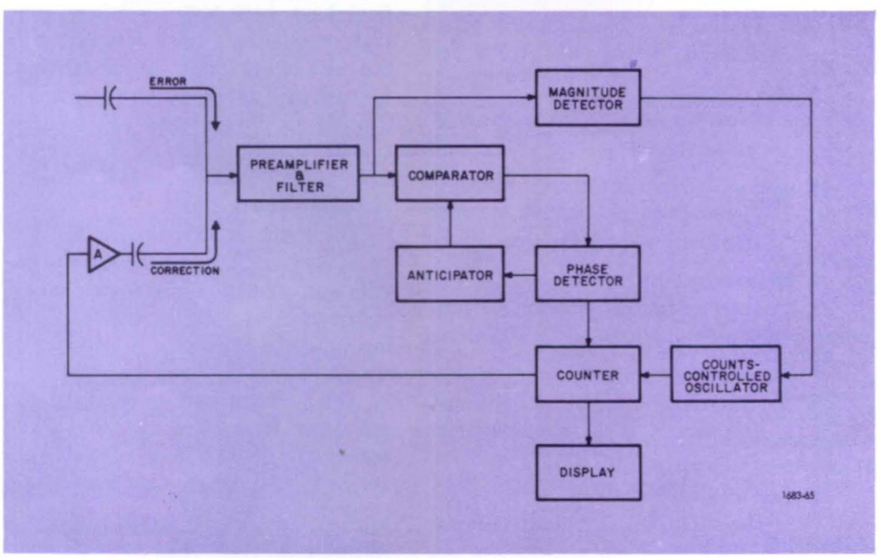


Figure 1. GR 1683 balancing technique.

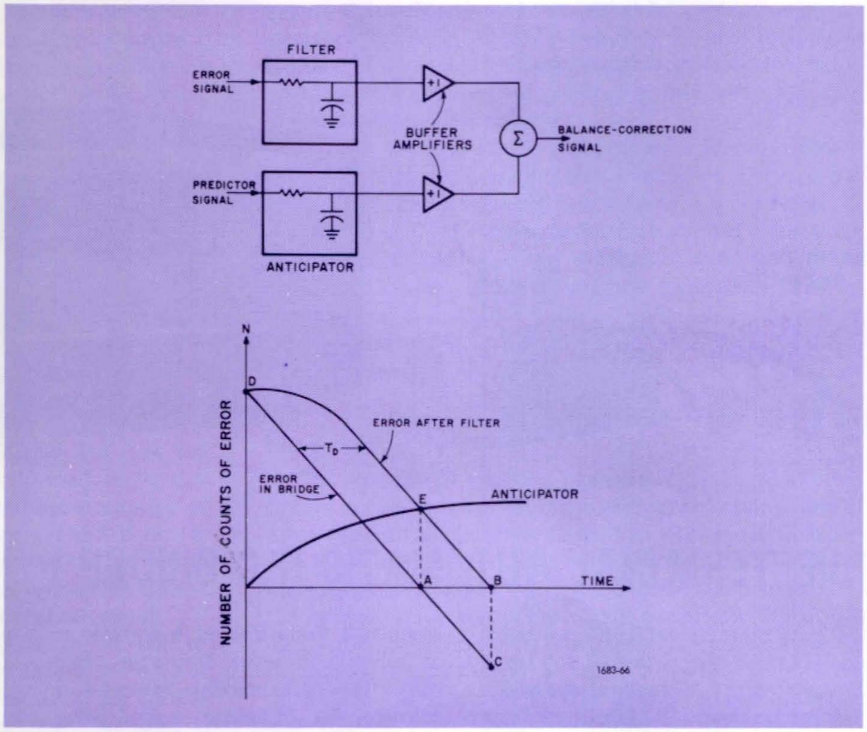


Figure 2. Anticipator operation.

that the bridge error signal is decreasing at a linear rate (line DAC). Because the filter introduces a time delay, we cannot detect the bridge error until some time (T_D) later (curve DEB). If we observe when this error reaches zero (point B), the actual bridge error has overshot and is now at point C. If, however, the error is detected at point E, by comparison of the error signal with a "predicting" function signal, the actual bridge error is zero and the bridge is balanced. The time necessary to recover from overshoot is saved.

The third time-saver is a bonus for those who are measuring a group of components with near-equal values. The GR 1683 starts a new balance from the point where the previous balance occurred. The closer in value the two components, the shorter the measurement time. Identical components require no balance time; a short start pulse is the only delay.

Features

- **Wide ranges are automatic**
  0.01 pF to 199.99 mF
  0.1 nH to 1999.9 H
  0.001 mΩ to 1999.9 kΩ
- **Standards**
  Uses stable, established GR standards of *C* and *R*.
- **Low measurement error**
  0.1% at the end of four feet of cable
- **Rapid speed of balance**
  125 ms at 1000 Hz for full-scale change; 1.3 s for 120 Hz for full-scale change; considerably faster for less-than-full scale changes
- **Bias**
  0 to 3 volts internal, up to 600 volts external
- **Leakage current detector**
  Measures leakage current and provides a go/no-go indication
- **Programmability**
  Most bridge functions are remotely programmable.
- **Data output**
  Various data-output options; compatible with integrated circuits
- **Test fixture**
  Available for axial-lead components

ABOUT THE BRIDGE *

Measuring capacitors, inductors, and resistors requires the efficient and economical use of bridge circuitry. The GR 1683 is an *active* bridge, as can be seen in Figure 3. The active elements are multistage, solid-state feedback amplifiers, with parameters an order of magnitude more stable than is required for bridge accuracy. Because these amplifiers have very high open-loop gain, their transfer function is determined by the ratio of the stable passive standards that are used as feedback elements. It is possible to obtain stable and accurate transfer functions to 0.005% or better. The bridge standards are GR precision capacitors and resistors.

The inductance/resistance and the capacitance bridges use the same amplifiers and standards interconnected accordingly. Equivalent-series-resistance measurement capability is achieved by the addition of one more amplifier.

The impedance of the current detector ranges from 10 mΩ to 100 Ω. This means that there will be a 0.01% error for each 160 pF of stray capacitance

*On page 19 you can read about the bridge we announced 37 years ago.

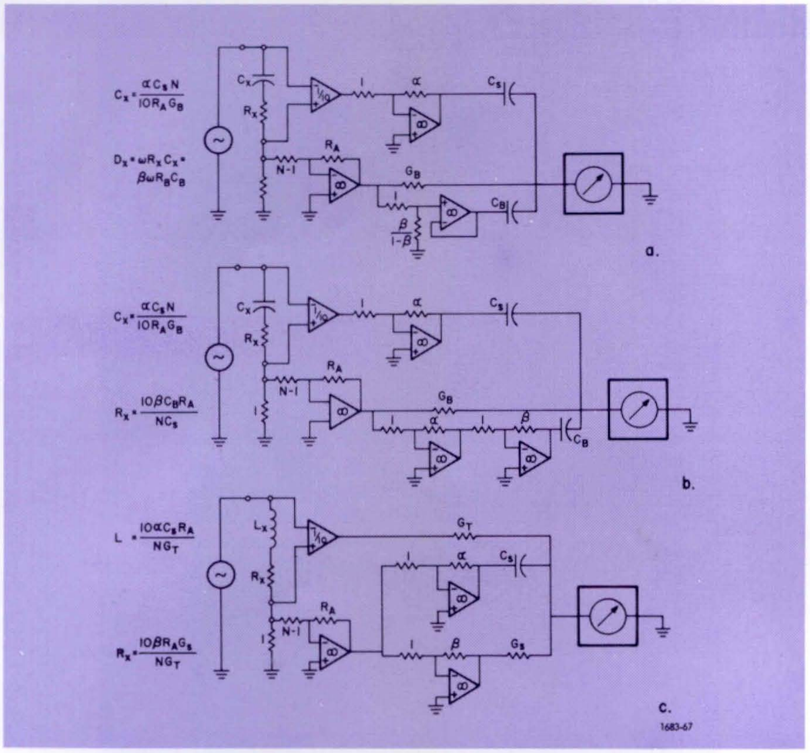


Figure 3. Simplified bridge circuits.

across the current-sensing circuit, when the measurement is made with the 100-Ω detector and at a frequency of 1000 Hz. This stray capacitance, moreover, causes an error in the quadrature signal (*D* measurement) rather than in the in-phase signal (*C* measurement), since the undesired current in the stray capacitance on the low side is 90° out of phase with the current into the resistive-current sensor and is usually negligible.

The wide range (mΩ to MΩ) of the bridge dictates the characteristics of the input amplifiers. For instance, accurate *low-impedance* measurements require a four-terminal bridge. Two terminals are used to inject and measure current through the unknown, and two terminals to measure the voltage across the unknown. In addition, accurate measurements of higher impedance require a low-impedance source to reduce the effects of stray capacitance, as well as to maintain bridge sensitivity during the measurement of low impedances. To satisfy these constraints, a low-output-impedance signal source and a low-impedance resistance ladder function as source and detector of the current through the unknown impedance.

A high-input-impedance differential amplifier is used to monitor the voltage across the unknown, making possible four-terminal measurements. In order to reduce stray-capacitance effects to near-zero, a fifth (ground) terminal is provided for a guard circuit.[2] The other amplifiers are precision voltage dividers that supply the current necessary to effect the balance.

[2] See page 16 for a description of the guarding technique used.

THE FINAL TEST

The degree to which *your* suggestion for improvements has been incorporated within the GR 1683 will be determined at *your* test benches. We look forward to a vote of approval.

– T. J. Coughlin

The GR 1683 was designed by the author; design and technical support were provided by D. S. Nixon, Jr. and A. W. Winterhalter. W. A. Montague was responsible for the mechanical design.

Complete specifications for the GR 1683 are in Catalog U.

Description		Price in USA
1683 Automatic RLC Bridge		
Bench Model		**$4250.00**
Rack Model		**4215.00**
Option 2 Remote Programmability	**add**	**200.00**
Option 3 Leakage Current	**add**	**100.00**
Option 4 ESR Readout	**add**	**225.00**
*Option 5A Low-Level Data Output	**add**	**200.00**
*Option 5B High-Level Data Output	**add**	**200.00**
1683-P1 Test Fixture for axial leads		**165.00**

*Not available together in the same instrument.

PATENT APPLIED FOR

Prices subject to quantity discount.

WHY MAKE A FIVE-TERMINAL BRIDGE?

Because we want to measure impedances in the milliohm to the megohm ranges accurately. Let's consider two extremes: two-terminal measurements of a 100-mΩ resistor and a 10-pF capacitor. Lead impedances would seriously affect measurement of the resistor; stray capacitances would make measurement of the capacitor fairly meaningless. Our approach to the design of the GR 1683 Automatic RLC Bridge* included careful consideration of the methods used to connect the component under test to the bridge.

THE STARTING POINT

Figure 1 demonstrates the effect of connection leads several feet long upon the measurement accuracy of a 100-mΩ resistor. A measurement error of 100% is clearly unsatisfactory. There are several solutions to the problem. A common method is that of substitution: the bridge terminals are shorted and the bridge balance is recorded. Then the unknown unit is connected to the terminals and another reading is recorded. The difference in readings is a measure of the unknown. This method, unfortunately, suffers from several uncertainties such as how close to zero resistance is the short circuit and how repeatable is the contact resistance? A better solution is a four-terminal measurement.

Four-Terminal Measurements of Low Impedances

From the data on Figure 2, such measurements appear to be perfect – no error! This is not completely true, however, for there are several potential sources of error which will increase measurement inaccuracy. A major error source could be a voltmeter of non-ideal characteristics. If the voltmeter input impedance is 1 MΩ and the voltage-terminal lead impedance is 100 Ω, the resultant error is 0.01%.

Another source of error is the lack of ability of the voltmeter to reject common-mode voltages, particularly if the voltmeter is grounded and has a differential input. This is demonstrated in Figure 3, in which a common-mode gain (rejection) of 10^{-6} is assumed. As

*See page 11.

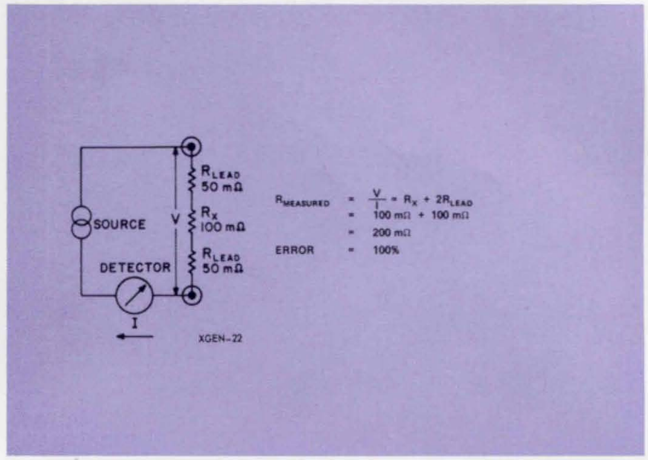


Figure 1. Effect of lead impedance on accuracy of resistance measurement.

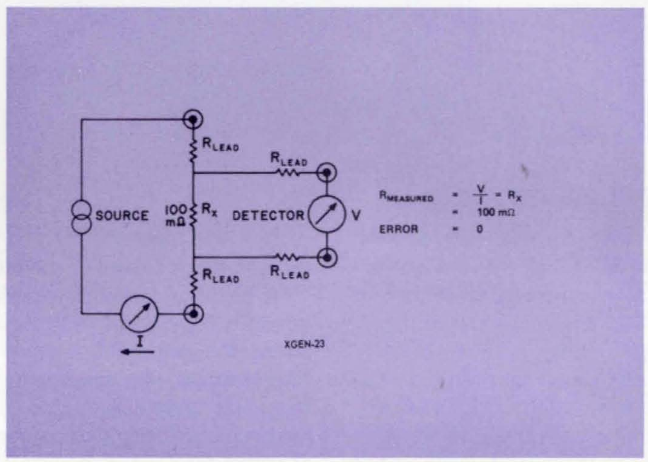


Figure 2. Four-terminal resistance measurement.

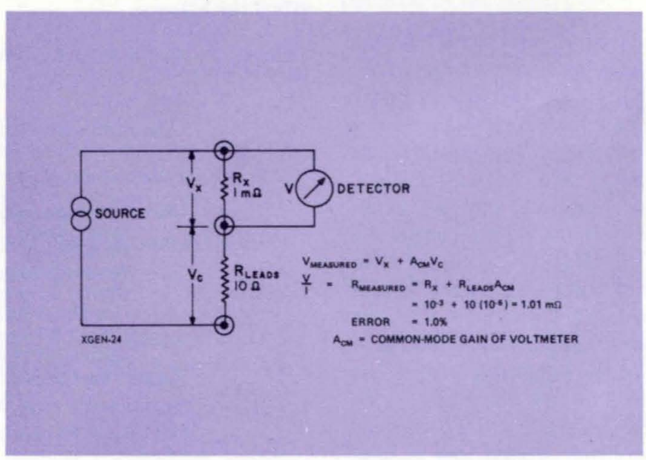


Figure 3. Effect of non-infinite common-mode rejection.

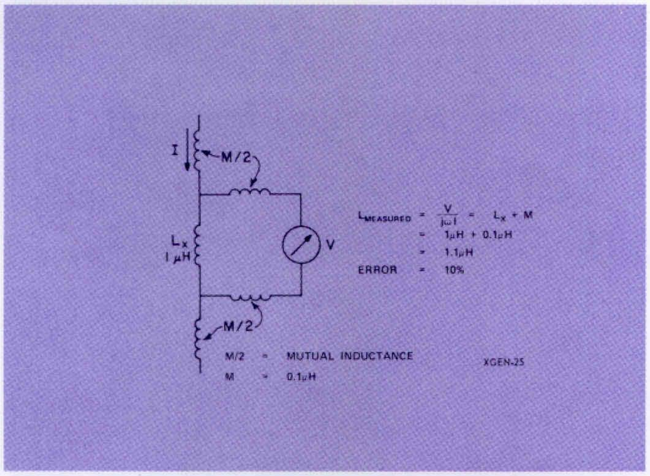


Figure 4. Degradation of inductance measurement by mutual inductance.

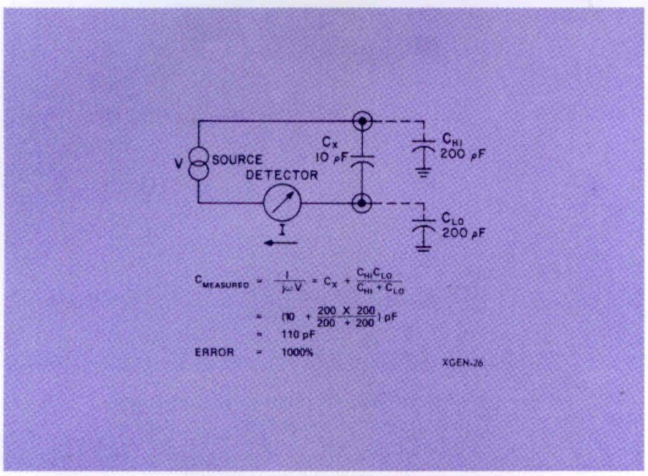


Figure 5. Two-terminal measurement of capacitance.

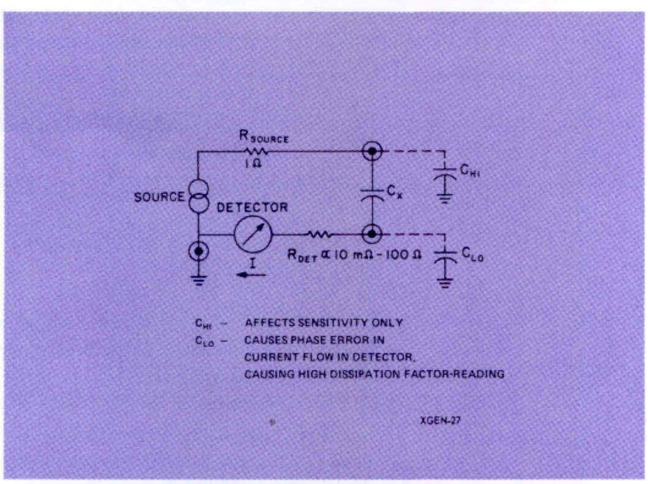


Figure 6. Three-terminal measurement of capacitance.

shown, the measurement error is 1%. Whether this size error is important is, of course, a function of the desired measurement accuracy.

There are more subtle errors in low-impedance measurements. For instance, consider mutual inductance between the current and potential leads.[1] The field induced by the current leads couples to the potential leads, thereby inducing an error voltage. Figure 4 is an example of the way in which an inductor-measurement circuit can be degraded. The solution is simple, if applied logically. Twisting together the current leads reduces the mutual induction by reducing the area of the enclosed field loop. A further improvement is obtained if the potential leads are twisted also and maintained at an angle of 90° from the current leads. If the 90° separation cannot be established, and the twisted pairs are nearly parallel, care must be taken that the pitches of the twisted leads are not made equal. The induced voltage will not be cancelled if the winding pitches are equal.

Three-Terminal Measurements of High Impedances

As mentioned previously, at the higher end of the impedance spectrum stray capacitance, not lead impedance, is the primary source of error.[2] Consider the two-terminal measurement of a 10-pF capacitor, as shown in Figure 5. Nominal values of stray capacitances have resulted in a measurement error of 1000%!

There are several remedies for this problem, one of which is the substitution method described earlier. The extra measurement is made, however, with the C_x circuit opened, which may affect the stray capacitances. A second, more preferred way, is to make a three-terminal measurement in which the effect of the stray capacitance is kept to a minimum by use of low-impedance circuitry, with the common source and detector point grounded to form a guard circuit. This is shown in Figure 6.

[1] Hall, H. P., "The Measurement of Electrolytic Capacitors," *GR Experimenter,* June 1966.

[2] Hersh, J. F., "A Close Look at Connection Errors in Capacitance Measurements," *GR Experimenter,* July 1959.

As the impedance of the source or detector circuitry increases, the effect of the stray capacitances on the accuracy of the measurement increases. The most usual form of this guarding technique occurs in a ratio-transformer capacitance bridge (Figure 7). In this case, stray capacitance C_{HI} is guarded by the low output impedance of the transformer, while the effect of C_{LO} is eliminated by the fact that at null there is no voltage across the detector terminals and, therefore, no current flows through C_{LO}.

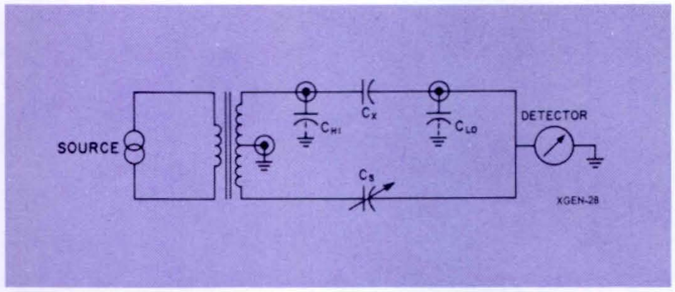


Figure 7. Three-terminal ratio-transformer bridge circuit.

THE WHY

We have had several requests for unusual measurements; one was to measure 10,000 pF at the end of 100 feet of cable. Our solution to this problem was a *five-terminal* measurement (Figure 8) made with the GR 1683 Automatic RLC Bridge. The four bridge terminals were connected to the unknown terminals to reduce the effects of the lead impedances to a minimum, and a fifth terminal (or guard) was connected to a common point to minimize the effects of the terminal capacitances. The bridge read C_x to ±1 count both at the bridge terminals and 100 feet away. The *D* reading increased by 0.0040 at 100 feet. This was extremely close to the calculated value. In this instance, a two-terminal measurement would have sufficed and the resultant *D* error would have been +0.0020, which was the measured value. The circuit error analysis is shown in Table 1.

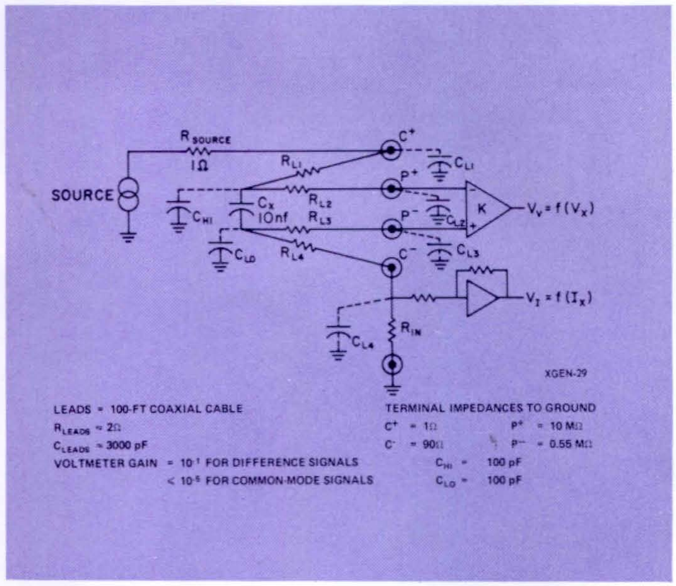


Figure 8. Five-terminal measurement of capacitor 100 feet from GR 1683.

Another request was to measure a 1-Ω resistor 25 feet away from the bridge (Figure 9). The reading at the end of the wire was in error by two counts, the calculated error plus the bridge uncertainty (±1 count). Most of the phase error due to the mutual inductance of the leads was eliminated by the twisting of the current leads. The circuit error analysis is shown in Table 2.

Table 1
Error Sources and Resultant Effects for Measurement of 10 nF

Source	Effect %	Effect counts	Comments
C_{L1}	0	0	Only affects sensitivity 160μF reduces sensitivity by 50%
R_{L1}	0	0	Only affects sensitivity
C_{HI}	0	0	
R_{L2}	0	0	$R_{L2}C_{L2}$ combined cause phase shift of 0.005%, which produces ½-count error in the quadrature measurement (D)
C_{L2}	0.005	0.5	
R_{L3}	0	0	Same effect as $R_{L2}C_{L2}$
C_{L3}	0.005	0.5	
C_{LO}	0.005	0.5	Adds ½ count to D reading
C_{L4}	0.19	19	Adds 19 counts to D reading
C_{L3}	0.19	19	Adds 19 counts to D reading
R_{L4}	0	0	$R_{L4}R_{IN}(C^-)$ combined affect sensitivity through common-mode gain of amplifier K. In this case, the effect is negligible.
$R_{IN}(C^-)$	0	0	

CONCLUSIONS

A wide-range impedance bridge such as the GR 1683 requires three- and four-terminal capabilities. If we apply principles of the new math, adding *three-terminal* capability for higher-impedance measurements to *four-terminal* capability for the lower impedance measurements results in a *five-*

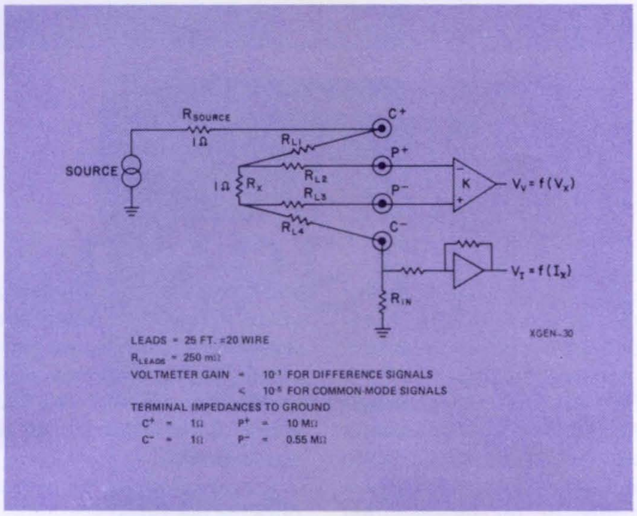


Figure 9. Five-terminal measurement of resistor 25 feet from GR 1683.

Table 2
Error Sources and Resultant Effects for Measurement of 1Ω

Source	Effect %	Effect counts	Comments
R_{L1}	0	0	Only affects sensitivity
R_{L2}	0	0	Negligible compared to voltmeter input
R_{L3}	0	0	
R_{L4}	0.0025	0.25	Effect due to common-mode gain of voltage detector K
$R_{IN}(C^-)$	0.01	1	

terminal bridge. If both stray capacitances and lead impedances affect the accuracy of the measurement, a five-terminal measurement will be required.

– T. J. Coughlin

T. J. Coughlin graduated from Northeastern University with the degrees of BSEE (1965) and MSEE (1967). He was a co-op student at General Radio in 1961 and joined GR in 1966 as a development engineer in the Impedance Group, specializing in design of pertinent instrumentation. He is a member of IEEE.

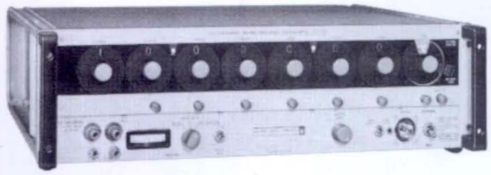
General Radio Expands

On March 3, 1970, details were completed for the purchase by General Radio of a controlling interest in Time/Data Corporation of Palo Alto, California. Time/Data specializes in development and production of high-speed electronic signal-processing instruments. It is the first domestic subsidiary to be acquired by General Radio in the 55-year history of the company.

Time/Data Corporation, in its four years of operation, has attained a recognized position in its special field of digital signal analysis. T/D's second-generation signal-processing device, the T/D Fast Fourier Transform Processor, was introduced in November 1969 at the Fall Joint Computer Conference. It is designed primarily for high-speed time-series analysis and synthesis under the control of an accessory computer. This permits analysis of electrical signals in real time with a speed and economy not possible with a computer alone. Such systems now are in use in oceanography, biomedical and geophysical research, radar signal processing, speech analysis, environmental science studies, analysis of medical data, and for structural-dynamics investigations that may include the analysis of vibratory characteristics of all types of products.

An announcement on March 3, 1970, designated D. B. Sinclair, GR's President, as President of Time/Data. L. J. Chamberlain has been transferred from General Radio, West Concord to assume the position of Executive Vice President, and E. A. Sloane has been named Vice President and Technical Director.

A second expansion move was made by GR on April 27, 1970, when officials of Grason-Stadler and General Radio signed agreements leading to purchase of all the stock of Grason-Stadler by GR. Grason-Stadler will operate within GR as a wholly-owned subsidiary, under its existing management. Grason-Stadler has been in business about twenty years. It has built a reputation as a leader in the commercial manufacture of precision audiometers and instruments for psycho-acoustics and the life sciences. Its experience and knowledge in these fields mesh with those of GR in the acoustics and signal-processing fields. The combined capability of the two companies is expected to exceed their individual capabilities by a wide margin. In particular, the marketing strength of GR is expected to contribute immediately to increased sales of G-S products. The over-all position of GR as a major source of acoustic instrumentation in the United States is expected to extend and solidify, particularly when taken in conjunction with the high technology of Time/Data's contribution in signal processing.



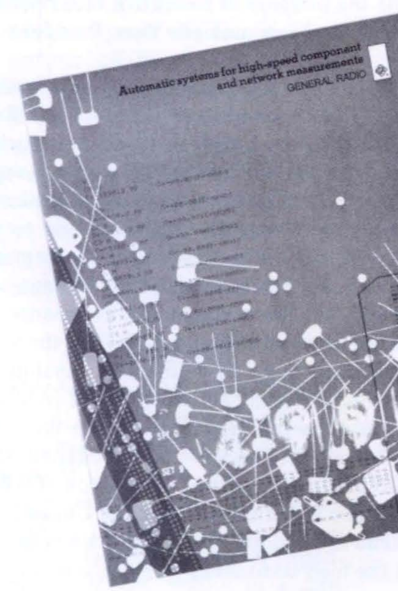
INFLATION DEFLATION

Price of GR instrumentation, as most readers know, is determined in advance of production release and is usually based upon several factors, one of which is anticipated sales. When initial development and other set-up costs have been written off, we are able to consider more favorable prices to the customers. At the same time, changes in design, to reduce manufacturing costs, are considered, but only if there is no reduction in performance.

Customer acceptance of the General Radio line of frequency synthesizers has been most gratifying. Design and initial setup costs have been written off, and the redesign of some modules has resulted in reduced manufacturing costs but, as mentioned above, with no reduction in performance. As a matter of fact, units that are frequency programmable have had switching time reduced to 200 ms. As a consequence, prices for some models of the GR 1161, 1162, 1163, and 1164 series have been substantially reduced. Readers who have need for synthesizers in applications such as Nuclear Magnetic Resonance, Ultrasonic Studies, Communication Oscillators, Crystal-Filter and Acoustic Measurements, or for incorporation into production test consoles, should review the table of price savings below. They should note, also, that the modular construction of the synthesizer units permits tailor-made resolution to fit the requirements of specific applications, simply by omitting unnecessary modules. Prices for reduced-resolution synthesizers, for instance, can be computed by the deduction of $230 for each decade removed and $300 for removal of the continuously adjustable decade (CAD).

Listed below are the basic specifications for the 7-digit synthesizers with a CAD. Complete specifications and new prices of all models are contained in the catalog pamphlet "Frequency Synthesizers," available upon request to General Radio or at the nearest GR District Office.

	1161-AR7C Old	1161-AR7C New	1162-AR7C Old	1162-AR7C New	1163-AR7C Old	1163-AR7C New	1164-AR7C Old	1164-AR7C New
Price in USA	$6785	**$4990**	$6940	**$4990**	$6920	**$5290**	$7695	**$7195**
Frequency Range	0 – 100 kHz		0 – 1 MHz		30 Hz – 12 MHz		10 kHz – 70 MHz	
Resolution	0.01 Hz		0.1 Hz		1.0 Hz		10 Hz	



Information Retrieval

Readers can obtain an index for the 1969 issues of the *GR Experimenter* upon request to the Editor.

For those interested in automatic systems for processing capacitors, semiconductors, resistors, and inductors, a newly-published brochure describing GR's systems capabilities is available. Included in the brochure are systems for measuring *R*, *L*, *C*, dissipation factor, leakage current, and other parameters. The examples in the brochure cover bench-top set-ups to multi-bay systems; peripheral equipment such as mechanical handlers and sorters, conditioning chambers, and computers also are included. Most of the systems illustrated will test components and networks to the stringent requirements of military specifications. Your copy of "Automatic Systems for High-Speed Component and Networks Measurements" is free upon request to General Radio.

The GENERAL RADIO EXPERIMENTER

VOL. VII. Nos. 11 and 12 — APRIL - MAY, 1933

ELECTRICAL COMMUNICATIONS TECHNIQUE AND ITS APPLICATIONS IN ALLIED FIELDS

THE CONVENIENT MEASUREMENT OF C, R, AND L

THE important considerations in the large majority of bridge measurements made in the average experimental laboratory are the ease and speed of making the readings, and the ability to measure any values of resistance, inductance, or capacitance, as they may exist in any piece of equipment. A completely satisfactory bridge should immediately indicate the answer to such questions as the following:

Is the maximum inductance of this variable inductor at least 5 mh, its minimum inductance 130 μh, and its direct-current resistance less than 4 Ω?[1]

Has this choke coil at least 20 h inductance and an energy factor Q of at least 20?

Has this tuning condenser a maximum capacitance of 250 $\mu\mu$f and a 20 to 1 range?

Has this filter condenser at least 4 μf capacitance and a power factor of only 0.5%?

Is the resistance of this rheostat 200 kΩ?

Is the zero resistance of this decade-resistance box only 5 mΩ?

The TYPE 650-A Impedance Bridge will furnish the answers to all these questions and many others. It will measure direct-current resistance over 9 decades from 1 mΩ to 1 MΩ, inductance over 8 decades from 1 μh to 100 h, with an energy factor ($Q = \frac{\omega L}{R}$) up to 1000, capacitance over 8 decades from 1 $\mu\mu$f to 100 μf, with a dissipation factor ($D = R\omega C$) up to unity.[2]

These results are read directly from dials having approximately logarithmic scales similar to those used on slide rules. The position of the decimal point and the proper electrical unit are indicated by the positions of two selector switches. Thus the CRL multiplier switch in Figure 1 points to a combined multiplying factor and electrical unit of 1 μf so that the indicated ca-

[1] These are the standard abbreviations of the Institute of Radio Engineers. Note that 1 mΩ is 0.001 ohm and that 1 MΩ is 1,000,000 ohms.

[2] The fact that this bridge is capable of measuring a condenser with large energy losses makes it necessary to distinguish between its dissipation factor $\frac{R}{X}$ and power factor $\frac{R}{Z}$. The two are equivalent when the losses are low. Since the bridge measures $R\omega C$ directly, the term dissipation factor has been used, even though the two terms are, for most condensers, synonymous.

pacitance as shown on the CRL dial is 2.67 μf, because the D-Q multiplier switch has been set on C for the measurement of capacitance. It also shows that the DQ dial is to be read for dissipation factor D with a multiplying factor of 0.1 yielding 0.26.

If the condenser had a smaller dissipation factor, this D-Q multiplier switch would have been set for the D dial with a multiplying factor of 0.01. Thus the D dial, as shown in Figure 1, indicates a dissipation factor of 0.0196 or a power factor of 1.96%.

For the measurement of pure resistance the D-Q multiplier switch would be set at R so that the CRL dial indicates a resistance of 2.67 Ω.

For the measurement of inductance the D-Q multiplier switch would be set at L and the CRL dial indicates 2.67 mh. Using the DQ dial the multiplier is 1 and the energy factor Q as shown in Figure 1 is 2.6. Had the coil under measurement been a large iron-core choke coil, the CRL multiplier switch might have been set at the 10 h point, thus indicating 26.7 h. Then the D-Q multiplier switch would have been set to indicate the Q dial with a multiplier of 100 and an energy factor Q of 41 as read on the Q dial.

The ease of balancing the bridge depends on the use of the logarithmically tapered rheostats and the two multiplier switches. To illustrate this, take first the measurement of direct-current resistance.

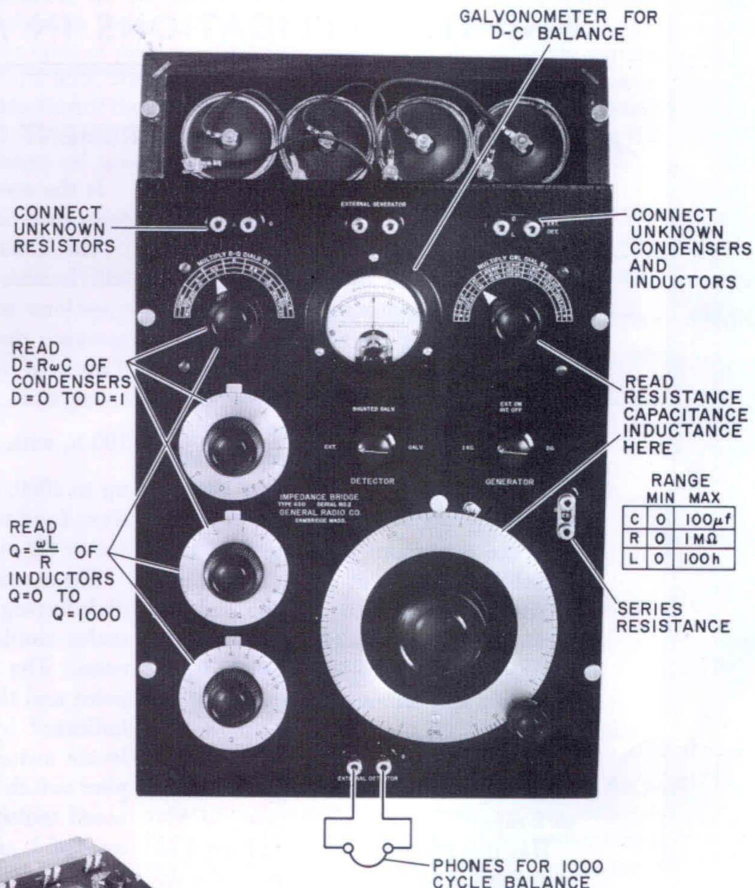
With the unknown resistor connected to the R terminals, the D-Q multiplier switch is set at R, the GENERATOR switch at DC, and the DETECTOR switch at SHUNTED GALV. The galvanometer immediately deflects, indicating by the direction of its deflection which way the CRL multiplier switch should be turned to obtain approximate balance. The CRL dial is then turned for exact balance, having thrown the DETECTOR switch to the GALV. position.

Because the calibration of the CRL dial extends to 0, the bridge can be balanced for a number of different settings of the CRL multiplier switch. This is very helpful in ascertaining the approximate value of a resistor. Obviously greatest accuracy of reading is obtained when the balance point on the CRL dial is within the main decade which occupies three-quarters of its scale length.

An inductor or condenser is measured by connecting it to the CL terminals. The GENERATOR switch is set at 1 KC. and the DETECTOR switch at EXT, head telephones being connected to the EXTERNAL DETECTOR terminals. The D-Q multiplier switch is set on L or C as the case demands, pointing to the DQ dial. The CRL dial is swept rapidly over its range to indicate the direction of balance. The CRL multiplier switch is then moved in the direction indicated and balance obtained on the CRL dial. The DQ dial is then turned for balance. From its setting the desirability of using the D dial or the necessity of using the Q dial will be indicated.

The reactance standards are mica condensers having all the excellent characteristics of the Type 505 Condensers described in the *Experimenter* for January.

The bridge circuit used for measuring condensers is the regular capacitance bridge having pure resistances for its ratio arms. Maxwell's bridge is used for inductors, whose energy factors Q are less than 10. Above this value Hay's bridge is used. The interdependence of the two balances of these last two



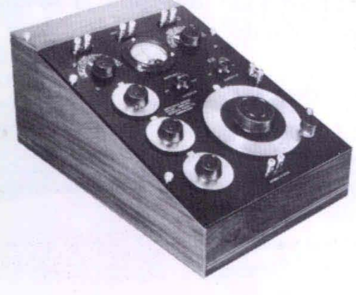


Figure 1. This photograph of the panel emphasizes the simplicity and wide range of the impedance bridge. In the corner at the left is a side view of the instrument.

bridge circuits cannot, of course, be prevented, but the use of the logarithmic rheostats for balancing makes it very easy to follow the drift of the balance points.

The accuracy of calibration of the CRL dial is 1% over its main decade. It may be set to 0.2% or a single wire for most settings of the CRL switch. The accuracy of readings for resistance and capacitance is 1%, for inductances 2%, for the middle decades. The accuracy falls off at small values because the smallest measurable quantities are 1 mΩ, 1 μμf, and 1 μh, respectively. Zero readings are approximately 10 mΩ, 4 μμf, and 0.1 μh respectively. The accuracy falls off at the large values, becoming 5% for resistance and capacitance and 10% for inductance. The accuracy of calibration of the DQ dials is 10%. The accuracy of readings for dissipation factor and energy factor is either 20% or 0.005, whichever is the larger.

The power for the bridge is drawn from four No. 6 dry cells mounted at the back of the cabinet. The liberal size of these batteries assures a very long life. External batteries of higher voltage may be used to increase the sensitivity of the bridge for the measurement of the highest resistances. The internal batteries operate a microphone hummer for the production of the 1-kc current. The capacitance of this hummer to ground is small and has been allowed for in the bridge calibration.

An external generator may be used, though its capacitance to ground may introduce considerable error. Subject to this limitation, the frequency may be varied over a wide range from a few cycles to 10 kc. The reading of the CRL dial is independent of frequency. The readings of the D and DQ dials must be multiplied by the ratio of the frequency used to 1 kc to give the correct values of dissipation and energy factors, while the reading of the Q dial must be divided by this ratio. For frequencies other than 1 kc the ranges of the DQ dials are altered so that they will no longer overlap. Additional resistance may be inserted by opening the SERIES RES. terminals. The Type 526 Rheostats, described on page 7, are quite satisfactory for this use.

– Robert F. Field

Design Assistance — Noise in Noise

Noise is a subject so much in the news these days we felt that a bit of information generated by Dr. Gordon R. Partridge, of General Radio, might benefit technical people engaged in noise-reduction efforts. This note simplifies and complements an earlier article in the *Experimenter.*[1]

A problem often encountered in sound measurements is the case of adding a new sound source to an existing background noise and finding the sound-pressure level that the new source *would have added* in the absence of the background. This note explains how to make the conversion from the measured values of background level and background level plus the added source.

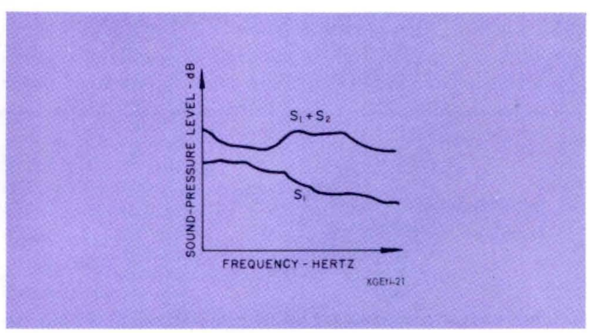
Call the background noise signal S_1 and the sum of the background level plus the added signal $S_1 + S_2$. Plot both of these measured signals (measured in one-third-octave bands, for instance) on the same graph, as shown in the sketch. Measure the difference in decibels between the $(S_1 + S_2)$ and the S_1 curves. This difference is tabulated as DIFF in the accompanying table. For the value of DIFF find the correction (CORR). *Subtract* this value of CORR from the $(S_1 + S_2)$ curve to find the decibel level of signal S_2 alone.

Example: The sound-pressure level in a room is 40.5 dB in the one-third-octave band centered at 315 Hz. An air-conditioner is turned on, and the sound level rises to 42.5 dB. What sound-pressure level does the air-conditioner alone produce?

Solution: The difference is 2.0 dB, so DIFF is equal to 2.0. The corresponding CORR is 4.3 dB. Subtract 4.3 from 42.5, obtaining 38.2 dB. Therefore, the air-conditioner by itself produces a sound-pressure level of 38.2 dB in the one-third-octave band centered at 315 Hz.

[1] Packard, L. E., "Background Noise Corrections in the Measurement of Machine Noise," *General Radio Experimenter,* December 1937.

DIFFerence	CORRection	DIFFerence	CORRection
0.1	16.4	12.0	0.3
0.2	13.5	13.0	0.2
0.3	11.8	14.0	0.2
0.4	10.6	15.0	0.1
0.5	9.6	16.0	0.1
0.6	8.9	17.0	0.09
0.7	8.3	18.0	0.07
0.8	7.7	19.0	0.06
0.9	7.3	20.0	0.04
1.0	6.9	22.0	0.03
2.0	4.3	24.0	0.02
3.0	3.0	26.0	0.010
4.0	2.2	28.0	0.006
5.0	1.7	30.0	0.004
6.0	1.3	32.0	0.002
7.0	1.0	34.0	0.001
8.0	0.7	36.0	0.001
9.0	0.6	38.0	0.000
10.0	0.5	40.0	0.000
11.0	0.4		



MORE MEMORY - MORE CONVENIENCE

Further development work on the Option 2 feature of the GR 1790 Logic-Circuit Analyzer[1] has resulted in an increase in the amount of data that can be stored, plus more convenience to the customer. The initial design of Option 2 incorporated a DEC disk with 32,000 words of storage. The improved form includes two cassette-type magnetic-tape transports, which are capable of storing more than 100,000 words per tape and of providing an unlimited program-tape library.

Ability to duplicate tapes is a feature not matched by the disk system. An additional feature of the improved option is its physical location within the GR 1790 console in the area previously occupied by two storage drawers.

The ease of preparing programs has been increased. To prepare test programs, type the text on the teletypewriter keyboard or read through the high-speed reader text previously punched in a paper tape. Test programs can be translated and executed without use of paper tape. Source and binary test programs can be stored, modified, or made accessible by commands typed on the teletypewriter. This feature promotes rapid program preparation and on-line modification of test programs.

The formal presentation of the features of Option 2 is: OPTION 2, EXTENDED MEMORY. Two cassette-loaded, magnetic-tape transports, one of which contains the system programs and the other the user programs. The cassettes contain 300 feet of certified tape capable of holding 100,000 words. Data-transfer rate is approximately 330 words per second (5 ips), and access speed is more than 6700 words per second (100 ips) in either direction. Time typically required for access to the entire 300 feet of tape is 30 seconds.

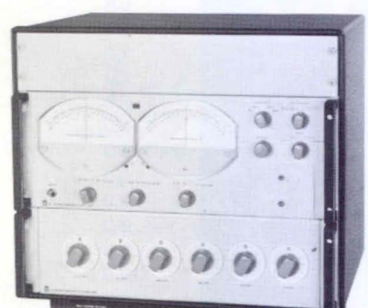
Access to any program on a tape is provided through software. Generation and execution of test programs are accomplished without the use of paper tape. Test programs of over 100,000 words in length are implemented with no support from the operator.

[1] "GR 1790 Logic-Circuit Analyzer," *GR Experimenter*, January/February 1970.

Description	Price in USA
1790 Logic-Circuit Analyzer, console version	**$32,500.00**
Option 1 Rack Version	(no extra charge)
Option 2 Additional Memory	**add 11,500.00**

Prices subject to quantity discount.

GR 1654-Z2 Sorting System



NEW

GR 1413 Precision Decade Capacitor

The GR 1413 Precision Decade Capacitor provides, in a single package, a capacitance standard that covers most widely used values. It was designed principally for use with the GR 1654 Impedance Comparator;[1] the combination is listed as the GR 1654-Z2 Sorting System. The capacitor ranges from 0 to 1.11111 μF in increments as small as 1 pF and with an accuracy of 0.05% + 0.5 pF. Air capacitors are used for the two lower decades and precision silvered-mica capacitors for the remainder. The unit is provided with three-terminal connections that can be altered easily to two-wire connections. Development was by R. W. Orr.

[1] Leong, R. K., "Impedance Comparison Sprints Ahead," *GR Experimenter*, May/June 1969.

Prices on page 23.

NEW

GR 1442 Coaxial Resistance Standard

Availability of the GR 1442-C/D/E Coaxial Resistance Standards, with values of 0.5, 1.0, and 2.0 Ω respectively, extends the range of GR coaxial resistors from 1,000 to 0.5 Ω. It is possible to establish very low values of standard dissipation factors for calibration of impedance bridges when these resistances are used in conjunction with the GR Coaxial Capacitance Standards.

Development of these standards was by R. W. Orr.

Prices on page 23.

GR 1656 Impedance Bridge

NEW

An inexpensive 0.1% CRL bridge with fast-balance lever switches, designated the GR 1656 Impedance Bridge.

The basic limitations on the accuracy of our 1% impedance bridge (GR 1650-B)[1] are the resolution and accuracy of the main rheostat and its dial. If a decade resistor were used, it would be relatively easy to tighten the tolerances on the internal standards to get a more accurate bridge. However, decade resistors (with the exception of the new GR 1436) have a row of knobs or concentric knobs which, while satisfactory for occasional adjustment, are tiresome for those who must continually balance bridges. Our most accurate universal bridge (the GR Type 1608,[2] now 0.05%) solves this problem by use of a special 100-position switch, so that only two concentric knobs are needed. This assembly, however, is comparatively expensive. Its use, plus installation of many other measurement and convenience features, results in a relatively expensive general-purpose instrument.

We think we've found the answer to the problem of designing a quickly balanced, high-resolution bridge. Our precision capacitance bridge (GR Type 1615)[3] uses a lever switch with digital readout which we, and our customers, have found very convenient. While this switch design is too expensive for a low-price instrument, we have, with the help of the Oak Manufacturing Company, developed a new lever switch for general-purpose use. This switch is the main readout control in the new GR 1656. With it, we have substantially reduced the time required to make a balance and greatly simplified the work required to set several digits to zero – with one sweep of the hand!

The improved readout resolution of the GR 1656, in addition to allowing better accuracy in readings, offers two other advantages. The smallest C, R, G or L that can be detected is extended by a factor of ten to 0.1 pF, 100 $\mu\Omega$, 100 p℧ (or pico-siemens) and 0.1 μH, respectively, and standards of even-decade values can be compared to 0.01%.

Another important improvement is the sensitive field-effect-transistor chopper-type dc detector that provides good sensitivity over all ranges of the R and G bridges, from 10^{-4} to 10^{10} Ω. This wide range, with its basic 0.1% accuracy, makes the GR 1656 a good dc resistance bridge as well as a good ac bridge.

In other respects the new instrument is very similar to the popular GR 1650-B, having in common its six bridges (series and parallel C and L, plus R and G), its battery operation, its internal signal source and detector, and its high D and Q accuracy. The 0.001-D accuracy is particularly important in a 0.1% bridge for, in many circuits, such a difference in D is just as important as a 0.1% difference in the value of the parameter.

The obvious use of the new bridge is in component measurement, particularly those components of tight tolerance which have come into wide use during the past few years. If the detector sensitivity is adjusted to cause a given meter deflection for a given percent unbalance, it may be used for rapid go/no-go measurements. It can be used also for a variety of measurements on networks and electrical devices.

Development of the GR 1656 was by H. P. Hall, who also contributed the foregoing material.

Complete specifications for the items below are in Catalog U.

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	Coaxial Resistance Standard	
1442-9702	**1442-C,** 0.5Ω	**80.00**
1442-9703	**1442-D,** 1.0Ω	**80.00**
1442-9704	**1442-E,** 2.0Ω	**80.00**
	1656 Impedance Bridge	
1656-9701	Portable Model	**700.00**
1656-9702	Rack Model	**735.00**
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	1654 Impedance Comparator	
1654-9700	Bench Model	**1300.00**
1654-9701	Rack Model	**1250.00**
	1654-Z2 Sorting System	
1654-9702	(with decade capacitor)	**2230.00**

Prices subject to quantity discount.

[1] Havener, C. D., "The Universal Impedance Bridge – New Face, New Features," *GR Experimenter*, May 1968.

[2] Hall, H. P., "A Precise, General-Purpose Impedance Bridge," *GR Experimenter*, March 1962.

[3] Hersh, J. F., "Accuracy, Precision, and Convenience for Capacitance Measurements," *GR Experimenter*, August/September 1962.

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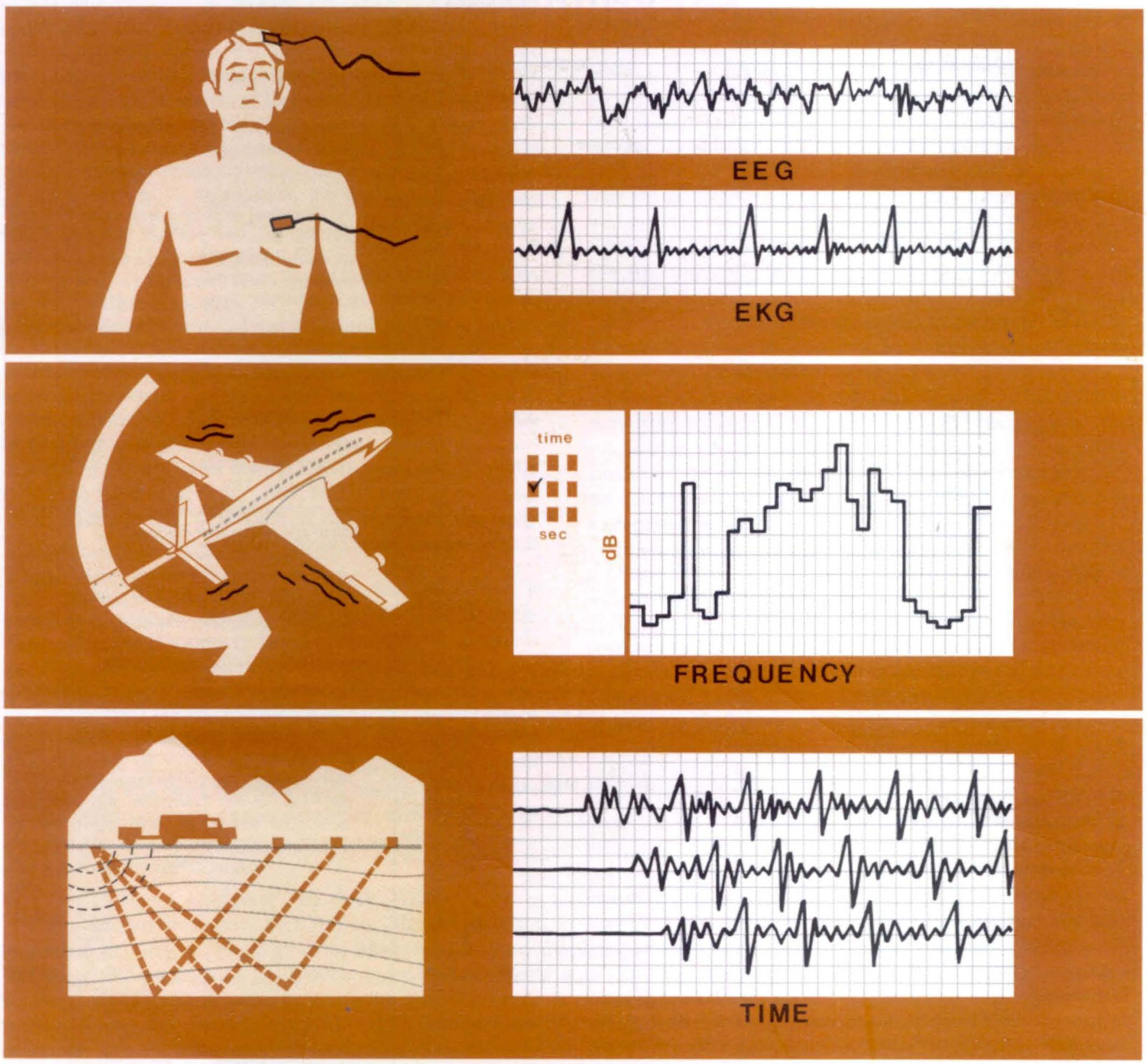

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The *General Radio Experimenter* is mailed without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General Radio Experimenter*, General Radio Co., Concord, Mass. 01742.

The Cover illustrates several of many successful applications of signal analysis in real time. The ability to match action with reaction, while events are taking place, is an excellent cost-reduction measure. It also permits knowledgeable experimental control during actions, as contrasted with hindsight decisions forced upon us by delayed data. The end result is an intelligent, economical approach to research, development, test, and production problems.

Complexity in the mathematical treatment of circuits and systems has been reduced in the past ten years. Small computers make possible our present-day electronic analyzers, programmed to apply all sorts of mathematical analytical techniques to design and test projects that will save engineering time and dollars. As part of simulation devices, the analyzers also can perform real-time functions in imitation of control by humans, thereby helping to reduce hazards to human operators.

The majority of design and test engineers are relatively too unskilled in the assembly and utilization of available instrumentation to create something which resembles, even remotely, the complex analytical test sets presently sold. This lack of skill is not in the mechanics of assembly but in the ***understanding of the techniques involved.***

There are, undoubtedly, numerous readers who bravely entered Norbert Wiener's classroom at M.I.T. to study the secrets of effective communication and of communication analysis. If, like this Editor, you departed from the seminar with a modicum of new knowledge and a plentitude of frustration, you will welcome another chance to tackle a phase of signal analysis, helped by the less difficult presentation on the next page. After that, perhaps, you will be ready for the age of the Fast Fourier Transform and associated techniques and for the instruments that incorporate them.

C E White

C. E. White
Editor

Signal Analysis with Digital Time-Series Analyzers

Although time-series analysis techniques are not new, they have not been used extensively because of the lack of suitable equipment, and therefore are not familiar to many engineers. The literature describing these techniques generally has been written for an academic audience and has been often clouded in abstract mathematics rather than being presented in user-oriented terms. This is an unnecessary obstacle to place before the potential user, because there is no more need to be a mathematician to use a time-series analyzer than there is to use a spectrum analyzer. In this article, the basic principles will be presented in non-rigorous physical terms as much as possible and will be illustrated with examples from several typical applications.

Analyzing electrical signals is a fundamental problem for engineers and scientists in all fields, whether they are working in the research and development laboratory, on the production line, or in the field. Whatever the physical, biological, or chemical system being studied, the basic phenomena can usually be converted into electrical signals by suitable transducers and analyzed to provide fundamental information about the system producing the signals.

Until recently, only a few basic analysis procedures were used, because of the limitations of the technology available for building test equipment. The basic signal-analysis procedure, other than direct observation of the signals on an oscilloscope, has been power-spectrum analysis, and will probably continue to be so. However, there are many other basic analysis procedures that would provide much more valuable information about the system under test, but they have not been available with the performance required for most practical problems. Even spectrum-analysis equipment, although serving adequately for a wide variety of problems, has severe limitations due to the analog-circuit technology available.

In the last decade, the availability of general-purpose computers allowed the advanced signal-analysis techniques that had been pioneered by Wiener, Lee, et al, to be developed to practical computational routines and were applied successfully to many data-analysis problems. They do not, however, help the experimenter or analyst who requires results rapidly so that he can interact with his system. The cost of computer analysis for this type of problem is also very high, especially when the loss of engineering time while awaiting results is taken into account.

The revolution in digital-processing technology that is now taking place has brought advanced signal-analysis techniques into the laboratory. This advance enables economic production of instrumentation systems which contain all of the computational power required for fast, accurate, signal processing. The TD 1923 Time-Series Analyzers, soon to be introduced by Time Data and GR, are the most advanced and complete line of this type of signal-processing equipment available. Besides their computational ability, they contain signal conditioning and display capability, with flexible, simple controls that allow them to be used as easily as conventional laboratory instruments.

WHERE AND WHY THEY'RE USED

Every process in nature gives rise to "signals" that are amenable to analysis by time-series analysis techniques. Therefore, the list of potential applications is endless. They can be grouped into categories based on the type of processing that is used and on the information that is desired. Some examples from several categories are illustrated on the front cover and described below.

Consider first some examples in the field of structural mechanics, where the basic quantities to be analyzed are mechanical vibrations in aircraft, automobiles, buildings, etc. These vibrations may be caused by many external forces such as wind, engine combustion and rotation, road roughness, earth tremors, and impact. The designer would like to determine such things as the source and transmission paths of the vibrations and the expected stresses and displacements at various points, by analyzing the signals from points on the structure. For instance, during the testing of an automobile, an objectional vibration may be found to exist in the passenger compartment. Time-series analysis techniques will determine if the vibration is coming from the engine, the road, or from the wind. In addition, the structural members that provide the transmission path for the vibration can be identified and suitably modified to filter out or to isolate the unwanted vibrations. The analysis system can even be used to simulate the structure and the external forces to test the new design before it is committed to production. Vibrations coming from the engine can also be analyzed to determine its condition. Failures may be predicted, or at least interpreted, from such a "signature analysis."

Another example is the determination of the flutter characteristics of aircraft in flight or in wind-tunnel tests. The resonant frequencies and damping of the various vibration modes of the aircraft are measured as a function of airspeed, to determine if there are any conditions that will produce excessive or unstable vibrations.

In the third example, rumbling vibrations, caused by automobiles driving across a bridge, are analyzed to pinpoint areas of high or unexpected loadings.

The field of biomedicine has already provided many important applications but it has only now begun to make use of sophisticated analysis techniques. The nervous system of the human body naturally produces electric signals whose characteristics are indicative of the condition of various parts of the body. The most common example is the electrocardiograph (EKG) signal, which provides information about the condition of the heart. Electroencephalograms (EEG's), or brainwave tracings, are analyzed to study brain damage and the effects of various stimuli and drugs on the brain. The techniques now being used by clinical physicians to analyze these signals are very primitive compared to those which can be used. The widespread use of time-series analysis techniques in medicine should increase as they become better understood.

Another application is geophysical exploration, an echo-ranging application that is in the same category as radar and sonar. A vibration signal is transmitted and echos are received from the reflecting geological strata. Comparison of the received and transmitted signals determines the time delay between them and, therefore, determines the location of the reflecting surfaces. Poor signals can be enhanced through analysis techniques, and a profile of the underground terrain obtained.

TIME-SERIES ANALYSIS

The preceding problems can all be solved with various combinations of the basic computational routines that constitute the tools of time-series analysis. By a time series or signal, we simply mean a succession of data values resulting from or simulating a physical process. The data can be amount of rainfall, earthquake tremors, electric voltages from a brain, or any other physical process. Time is usually the independent variable, but not always. The common thing about all these records is that the successive data values in each of the time series are related in some way to the other values in that time series, and perhaps to values in some other time series. These relationships may be deterministic or statistical. Time-series studies analyze these relationships so that the physical process can be better understood. This process can then be modeled, or simulated, either mathematically or physically.

The basic signal-analysis procedures required in a practical instrument can be grouped into three measurement categories, as shown in the table below. The processing functions must generally be fast enough so that data are processed in real time, that is, as fast as they are being acquired. Real-time processing also is necessary because the number of input data samples is usually much larger than the memory capacity of any practical machine. In a typical measurement, the input signal may be sampled for 10 minutes at a rate of 100,000 samples/second. The total number of input samples becomes astronomical but, after being processed by the appropriate analysis procedure, the data may have been reduced to only 100 numbers.

Basic Signal-Analysis Procedures

A. Measurement of Similarity
   1. Correlation analysis
   2. Spectral analysis
   3. Filtering

B. Measurement of Waveforms
   1. Ensemble averaging

C. Measurement of Statistical Distributions
   1. Probability density functions (amplitude distribution)

The product of the processing operation is often the final result desired. Frequently, however, some further processing must be performed on this result to put it into the form most suitable for showing the information being sought. Other processing functions allow this by performing basic arithmetic operations, coordinate transformations, smoothing operations, and time/frequency transformations.

MEASUREMENT OF SIMILARITY

The most important time-series analysis tools are those that give a measure of the similarity between signals. Spectral analysis, correlation analysis, and filtering provide this information. In fact, they provide the same measure of similarity, based upon the mean-squared difference between the signals, but present the results in different forms. Although spectral analysis and filtering are more familiar to most people than correlation, the latter offers more insight into the concept of measurement of similarity and therefore will be discussed first.

C. L. Heizman is a graduate of City College of New York (BEE - 1955) and Columbia University (MS - 1956). He was involved in research and product development of integrated circuits, high-speed computers, and data-analysis instruments prior to joining Time/Data Corporation (a GR Subsidiary) in 1968. As Applications Manager at T/D, Mr. Heizman devotes much time to real-time analysis. He is a member of IEEE.

Correlation Analysis

A natural way to compare the two waveforms of Figure 1 is to subtract one from the other, ordinate by ordinate, square each term to give quantities that are porportional only to the magnitude of the difference, and then to sum all the squared difference terms to obtain a single number that is a measure of the similarity. This number, when normalized by the number of independent measurements, is the mean-square difference. It can then be calculated for various displacements of one signal with respect to the other. By simple algebra, you can show the same information by calculating the correlation, or covariance, function which is the sum of the ordinate-by-ordinate multiplication of the two waveforms.* The result for the two random signals of Figure 1 shows that they are most similar when there is a displacement of 0.7 millisecond between them. If these signals represent the vibration level at two points on a structure, this time is the propagation time for vibrations between the two points. The sign of the displacement indicates the direction of propagation. Measuring time delays in this way leads to many useful applications.

Instead of calculating the cross-correlation function of two different signals as above, you can correlate a signal with itself to give the auto-correlation function. This shows how successive samples of a signal are related.

The auto-correlation functions in Figure 2, for the two random signals of Figure 1, show that successive samples of the upper signal of Figure 1 are more correlated, or are more dependent upon each other, than those of the lower signal.

Another example is shown, in Figure 3, of a sine-wave signal buried in noise. Even though the sine wave is not readily observable in the original signal, the auto-correlation function shows its presence clearly. This ability to detect periodic components in a signal is an important application of the auto-correlation function.

Spectral Analysis

The measurement of similarity is often easier to interpret when the operation is done as a spectral calculation. It is also

*The mean-square difference is

$$\epsilon_D = \frac{1}{N} \sum_n (x_n - y_{n-\tau})^2$$

$$= \frac{1}{N} \left(\sum_n x_n{}^2 + \sum_n y_n{}^2{}_{-\tau} - 2 \sum_n (x_n\, y_{n-\tau})^2 \right)$$

All the information about the similarity of the signals is contained in the third term, which is a maximum when the signals are most similar. This term is the correlation or covariance function.

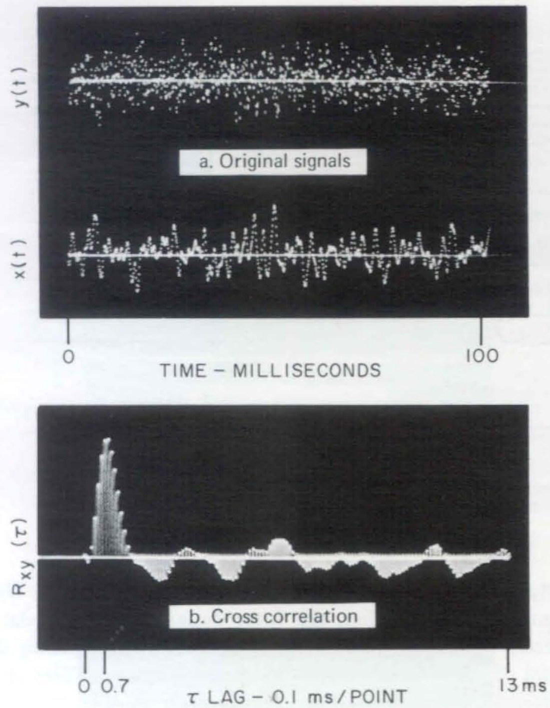


Figure 1. Cross-correlation function of two waveforms.

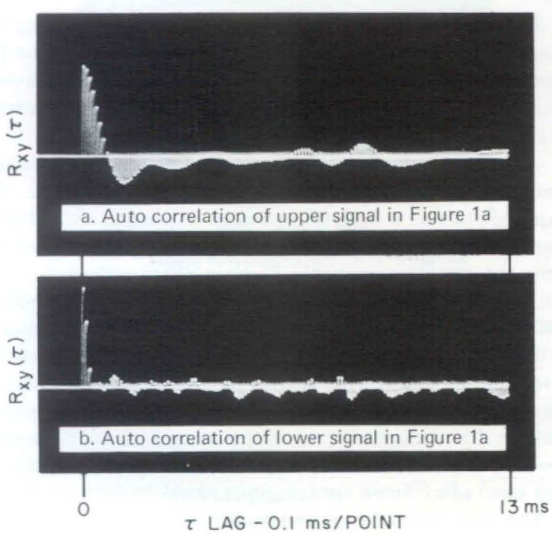


Figure 2. Auto-correlation functions.

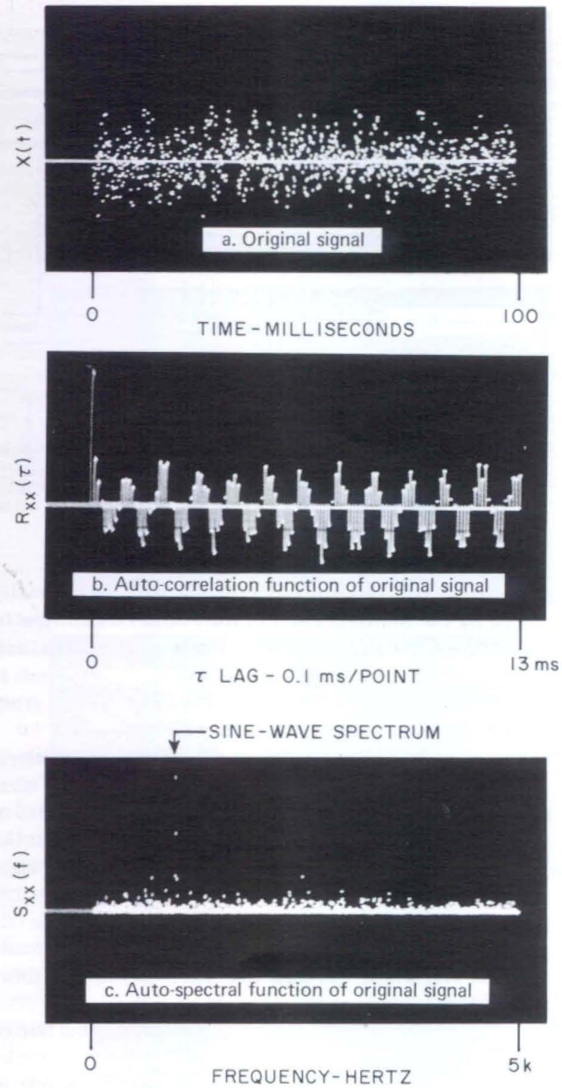


Figure 3. Analysis of sine wave buried in noise signal.

more common to see the measurements in this form because in the past spectral analysis was easier to implement with electronic instruments than correlation analysis.

Spectral analysis is based upon the theorem that any repetitive signal can be considered to be the sum of sinusoidal components whose frequencies are integral multiples of the basic repetition frequency. Fourier-transform analysis, familiar to most engineers and scientists, involves the calculation of the amplitudes and phase angles of these components. In principle, the calculation is really a cross correlation of the signal with a sinusoid of each of the possible harmonic frequencies, respectively. The result is always another sinusoid of the same frequency, whose amplitude and phase are proportional to the corresponding component in the signal. In practice, the Fourier-transform calculation is done in a much faster, more direct way by efficient computational methods that have been developed in recent years.

Auto-spectral analysis involves the calculation of the squared magnitude of the Fourier spectrum and is the quantity produced by most spectral analyzers. Because it is proportional to the power of a signal, it is commonly called Power Spectral Density (PSD). It gives exactly the same information as the auto-correlation function; in fact, it is the Fourier transform of the auto-correlation function, and it can be calculated in that way.

A spectral measurement that is not commonly available from analog spectrum analyzers, but which is extremely use-

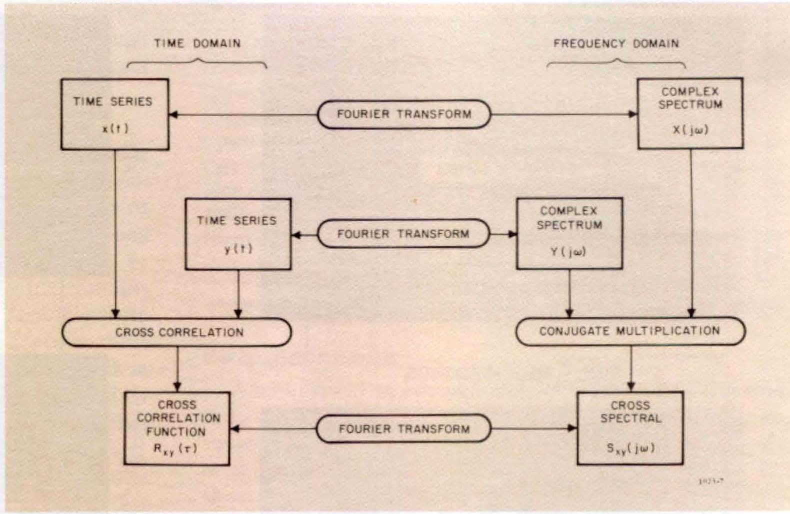


Figure 4. Time/frequency-domain diagram.

ful, is the cross-spectral function. This function is normally calculated by the multiplication of the Fourier transforms of two signals. Like the auto-spectral function, it contains exactly the same information about the similarity of the signals as the cross-correlation function, and it can be calculated from the cross-correlation by the Fourier transform.

The relation between correlation and spectral functions is illustrated by the time- and frequency-domain map of Figure 4. It shows that the correlation function can be calculated in two ways: directly in the time domain, or indirectly in the frequency domain via Fourier transforms. The fast algorithms developed to calculate Fourier transforms have enabled the calculations to be done more quickly via the frequency domain than in the time domain and thus have made the Fourier transform the key operation in modern time-series analyzers.

The auto-spectral function for the sine-wave signal buried in noise is also shown in Figure 3, along with the auto-correlation function. The sine-wave signal is clearly discernible in either function.

Filtering

A filter can be considered as a device that continuously compares an input signal with a stored reference signal and produces a maximum output when the two are most similar. The stored reference signal is the impulse response of the filter, i.e., the response of the filter to a very narrow pulse applied at the input. The criterion for the measurement of similarity is the same as for correlation and spectral analysis; in fact, the calculations are carried out in exactly the same way. Figure 5 shows a filter that was designed to detect the presence of a "chirp" or swept-frequency sine-wave signal buried in noise. This type of signal is used often as the transmitted signal in radar, sonar, and geophysical echo-ranging systems. The output of the filter increases significantly when the signal is present and has the same shape as the auto-correlation function of the signal.

The implementation of filters to "match" a prescribed signal is quite difficult with analog components, and practically impossible if the filter is to be easily variable. With digital implementation, the filter design becomes trivial – merely the specification in either the time or frequency domain of the waveform to be detected – and the filter characteristic can be changed in microseconds.

MEASUREMENT OF WAVEFORMS

Ensemble averaging* is very useful for determining the shape of a signal that is obscured by random noise when the signal is repetitive or when its time of occurrence is known. This latter condition exists when a system is being stimulated in a controlled manner.

An example is the electroencephalograph (EEG) signal produced by the flashing of a light in a person's eye. The responses to this stimulation (evoked responses) are added or averaged together. The signal-to-noise ratio is increased because the signal components, being in phase with each other, will add linearly while the noise components, being random, will partially cancel each other and add at a rate proportional to the square root of the number of averages.

Ensemble averaging is also commonly applied to the measurement of correlation and spectral functions to improve the statistical accuracy of the measurement. Consider the auto-spectral measurement of a short segment (1000 samples in this case) of a filtered random noise signal, as shown in Figure 6. The auto-spectral function itself is also a random function. Averaging the auto-spectral measurements of successive segments of the signal reduces the statistical variations.

MEASUREMENT OF STATISTICAL DISTRIBUTIONS

The amplitude histogram is often the first measurement made in the analysis of random data. You determine it by dividing the amplitude range into many equally spaced levels and by counting the number of times the measured value of the signal is at each level. The histograms of the two random signals of Figure 1 are shown in Figure 7. Besides showing the

*A statistical average evaluated from the probability density of a random process.

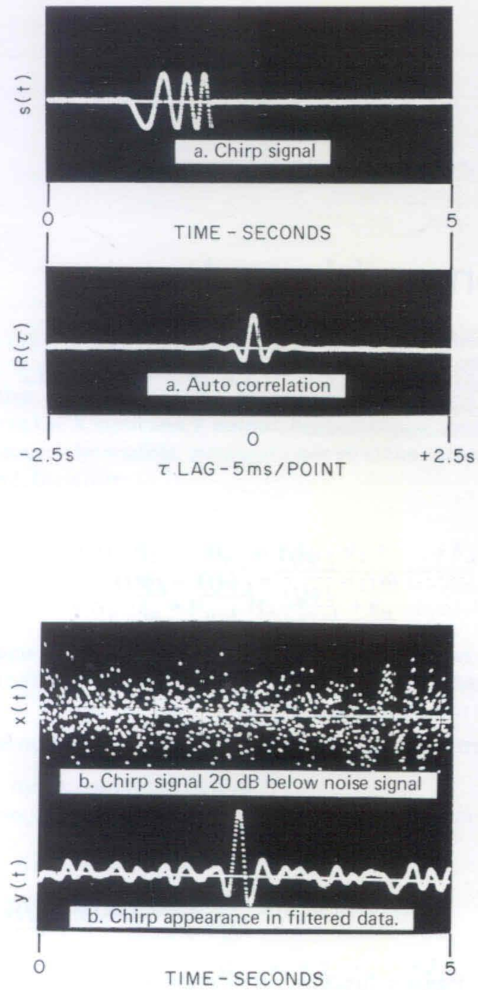


Figure 5. Signal detection by filtering.

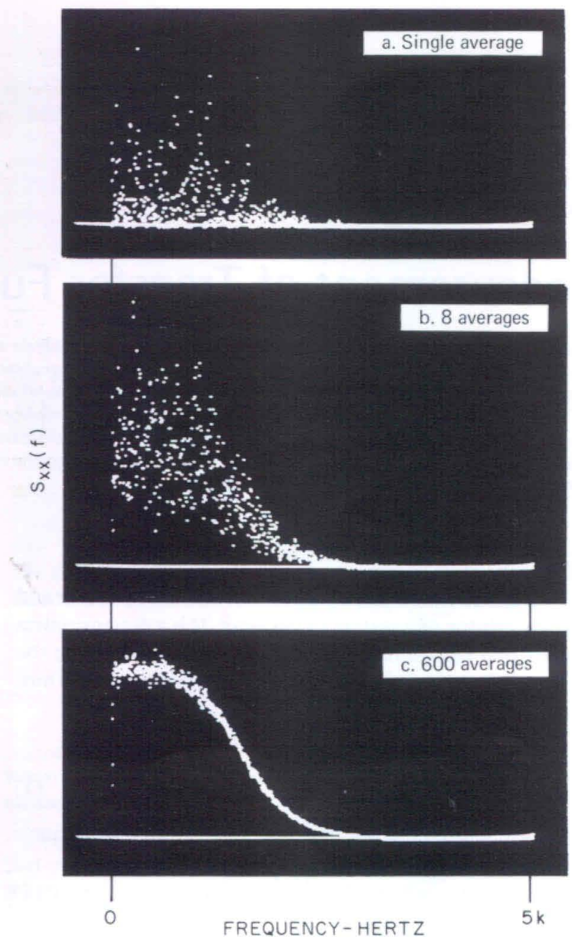


Figure 6. Averaged auto spectrum of band-limited noise.

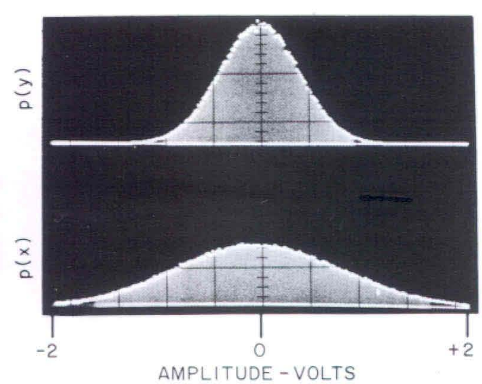


Figure 7. Amplitude histograms of random noise.

extremes of signal amplitudes to be expected, the amplitude distributions give information about the linearity of the relationship between the signals. If these signals represent the input and output of a system under test, the fact that both signals have the same distribution, Gaussian in this case, indicates that the system is probably linear.

From amplitude histograms, the mean value, rms value, and higher-order moments can be calculated.

SUMMARY

The basic tools required for analyzing signals have been discussed, and some of their applications have been given. The Fourier transform is seen to be a fundamental calculation required in digital time-series analyzers for doing correlation and spectral analysis and filtering. Many other calculations often are required, such as coherence function, transfer function*, cepstrum, etc, but these are extensions of the fundamental procedures described and are obtained by further processing. It is now practical, by application of modern digital technology, to build instruments that can accomplish all these functions with the performance required for real-world problems.

C. L. Heizman

*See page 8.

Measurement of Transfer Function and Impedance

The concept of transfer function is widely employed in the analysis and synthesis of linear systems or networks. Conventional instrumentation is well suited to the problem when a simple noiseless single input, single output system is involved, particularly if the rate at which measurements must be made is low. If not, modern FFT (Fast Fourier Transform) techniques provide an efficient means for performing these measurements at lightning speed. This paper derives the best (least mean-squared) transfer-function and impedance estimator, which is shown to involve the ratio of the input-output cross-spectral to the input auto-spectral density.

General

The transfer function of a linear network is a popular and useful descriptor of a system or network. It is a dimensionless quantity that relates the input and output by specifying the gain (or attenuation) and phase shift at all frequencies. Thus, the transfer function of the system shown in Figure 1 is

$$H(f) \angle\theta(f) = \frac{Y(f) \angle\psi(f)}{X(f) \angle\phi(f)} \qquad (1)$$

where

$$H(f) = \frac{Y(f)}{X(f)} \qquad (2)$$

is the gain of the system for a frequency of fHz as measured by the ratio of the output function $Y(f)$ and the input driving function $X(f)$,

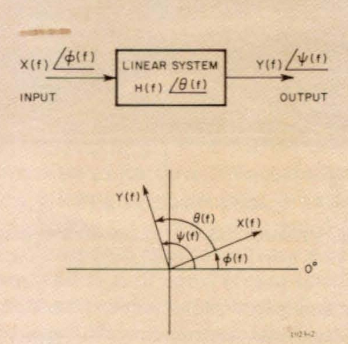


Figure 1. Linear-network derivation.

while

$$\theta(f) = \frac{\angle\psi(f)}{\angle\phi(f)} = \angle\psi(f) - \angle\phi(f) \qquad (3)$$

is the relative phase shift introduced by the system, as measured by the difference in phase angle between the input and the resultant output.

In terms of complex notation, the transfer function is

$$H(f)\exp[j\theta(f)] = \frac{Y(f)\exp[j\psi(f)]}{X(f)\exp[j\theta(f)]} \qquad (4)$$

$$H(f)\exp[j\theta(f)] = \frac{Y(f)}{X(f)} \exp\left(j[\psi(f) - \theta(f)]\right. \qquad (5)$$

The transfer function often must be measured for purposes of system analysis or simulation. Simulation is particularly important when the system under study is inaccessible or unwieldy for experimentation. Examples of this type might include simulation of process control systems or subsystems that can only be observed in operation, simulation of the response of space vehicles to transient excitation without running the risk of actual damage, or analyzing and simulating the effective transfer function of a vehicle suspension system. Knowledge of the transfer function would permit a convenient analog to be constructed.

Test equipment capable of providing the necessary measurements has been developed over the years and generally includes a sine-wave generator, voltmeters, and a phase-angle meter as the basic tools. The amplitude of the response and its relative phase angle constitute the necessary measurements. A measurement is usually obtained when the frequency of excitation is varied while the input is kept constant.

Standard measurement techniques of this type are usually adequate and are particularly suitable when the system under observation is

1) Noiseless
2) Has a single input port

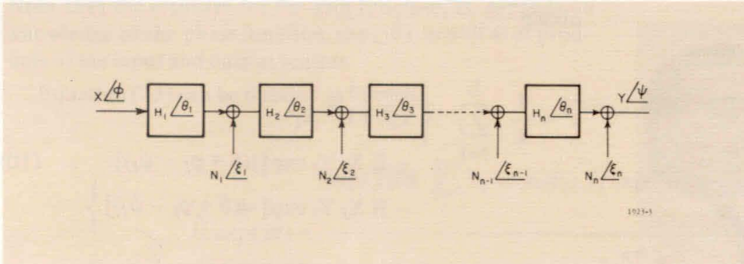


Figure 2. Noisy-network block diagram; multiple inputs.

In other words, if the system is as shown in Figure 2, the input $X\angle\phi$ and the output $Y\angle\psi$ are not simply related by the effective transfer function describing the gain and phase between the X input and Y output, because of the presence of numerous noise sources, propagating toward the output port. We find, therefore:

$$\begin{aligned} Y\angle\psi &= X H_1 H_2 \ldots H_n \angle \phi + \theta_1 + \theta_2 + \ldots + \theta_n \\ &+ N_n \angle \xi_n + N_{n-1} H_n \angle \xi_{n-1} + \theta_n \\ &+ N_{n-2} H_{n-1} H_n \angle \xi_{n-2} + \theta_{n-1} + \theta_n \ldots \\ &+ N_1 H_2 H_3 \ldots H_n \angle \xi_1 + \theta_2 + \theta_3 + \theta_n \end{aligned} \tag{6}$$

where the vector form $A\angle\alpha \equiv A(f)\angle\alpha(f)$.

Thus, the output $Y\angle\psi$ can be thought of as the sum of two vectors

$$Y\angle\psi = X\angle\phi\, H\angle\theta + N\angle\xi \tag{7}$$

where $H\angle\theta = H_1 H_2 H_3 \ldots H_n \angle \theta_1 + \theta_2 + \ldots + \theta_n$

and $N\angle\xi = N_n \angle\xi_n + N_{n-1} H_n \angle \xi_{n-1} + \theta_n + \ldots$

$$+ N_1 H_2 H_3 H_4 \ldots H_n \angle \xi_1 + \theta_2 + \theta_3 + \ldots + \theta_n$$

Thus, the system of Figure 2 can be replaced by an equivalent network as shown in Figure 3(a), provided that we are only concerned with finding the transfer function between the input terminals where $X\angle\phi$ is applied and the output terminals where $Y\angle\psi$ is observed. The vector relationships are shown in Figure 3(b).

It can be shown that a similar input-output relationship exists for any of the "n" input ports. Hence, if $H\angle\theta$ can be measured, "n" similar measurements will describe the total set of network transfer functions. In some cases, only one transfer function is required, that which applies between the X-input and Y-output terminals. Now, the signals designated as $N_1\angle\xi_1$ through $N_n\angle\xi_n$ can be thought of as internal noise sources that may not be directly observable. In either case, the problem is essentially the same: Measure the transfer function in the presence of other spurious signals, which we will conveniently call noise.

The equivalent noise vector has introduced uncertainty in the amplitude and phase measurements of the output vector $Y\angle\psi$. The measurement procedure must minimize these effects.

An analogous problem arises when the impedance of a noisy network is measured. In this case, the input vector of Figure 3 and equation (7), $Y\angle\psi$, becomes a voltage or velocity vector, $N\angle\xi$ a voltage or velocity noise vector, and $X\angle\phi$ the current or force vector. The equation relating the various factors is:

$$Y\angle\psi = X\angle\phi\, Z\angle\theta + N\angle\xi \tag{7a}$$

where $Z\angle\theta$ represents the impedance of the network to be measured.

With these differences in mind, the results and discussions that follow for estimating the transfer function will apply directly for estimation of the impedance function.

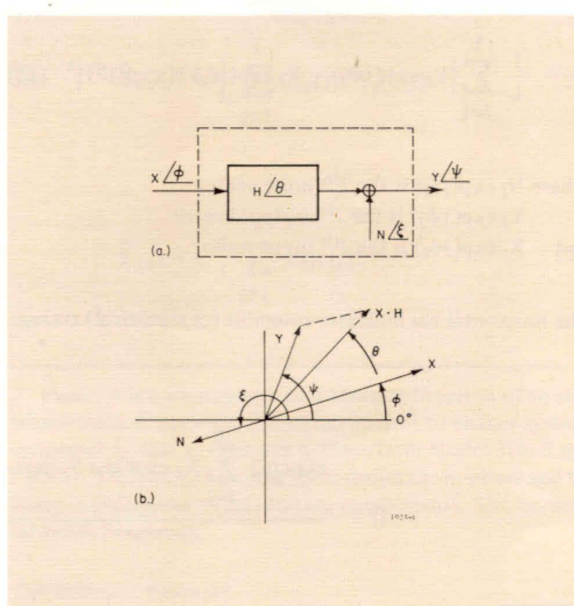


Figure 3. Noisy-network derivation.

E. A. Sloane received his BSEE degree from Northeastern University in 1951. He was a staff member at MIT Lincoln Laboratory, radar systems group, from 1951 to 1954; he joined Ampex Corp. as manager first for product planning, then for advanced technology planning, 1955-1965. Presently, he is Vice President and Technical Director of Time/Data Corporation, a GR subsidiary, which he helped found. Mr. Sloane is a member of Tau Beta Pi and Eta Kappa Nu.

Least-Mean-Square Estimation

Because the spurious signals are treated as a single equivalent noise vector, we can express it in terms of the transfer-function input and output as follows:

$$N\angle\xi = Y\angle\psi - X\angle\phi\ H\angle\theta \tag{8}$$

or $N\exp(j\xi) = Y\exp(j\psi) - X\exp(j\phi)\,H\exp(j\theta)$

The mean-squared value of this error after L* measurements will be:

$$\overline{\epsilon^2} = \frac{1}{L}\sum_{l=1}^{L}\left|N_l\exp(j\xi_l)\right|^2$$

$$= \frac{1}{L}\sum_{l=1}^{L}\left|Y_l\exp(j\psi_l) - X_l\exp(j\phi_l)\,H\exp(j\theta)\right|^2 \tag{9}$$

where $N_l\exp(j\xi_l)$ is the l^{th} noise vector
$Y_l\exp(j\psi_l)$ is the l^{th} output vector
and $X_l\exp(j\phi_l)$ is the l^{th} input vector.

The horizontal bar indicates ensemble (or statistical) average.

Performing the indicated operations in equation (9), we obtain

$$\overline{\epsilon^2} = \frac{1}{L}\sum_{l=1}^{L}\Big\{Y_l^2 + H^2X_l^2 - HX_lY_l\exp[j(\theta+\phi_l-\psi_l)] - HX_lY_l\exp[-j(\theta+\phi_l-\psi_l)]\Big\} \tag{10}$$

Because we desire to minimize this error by choosing an appropriate estimator** for the gain, $\hat{H}$, and the phase angle, $\hat{\theta}$, we proceed by differentiating with respect to the phase angle and setting the derivative to zero. Thus,

$$\frac{\delta\overline{\epsilon^2}}{\delta\hat{\theta}} = -\frac{1}{L}\sum_{l=1}^{L}H\Big\{X_lY_l\exp[j(\hat{\theta}+\phi_l-\psi_l)] - X_lY_l\exp[-j(\hat{\theta}+\phi_l-\psi_l)]\Big\} = 0 \tag{11}$$

so that

$$\exp(2j\hat{\theta}) = \frac{\sum_{l=1}^{L}X_lY_l\exp[-j(\phi_l-\psi_l)]}{\sum_{l=1}^{L}X_lY_l\exp[+j(\phi_l-\psi_l)]} \tag{12a}$$

$$\text{or}\quad \exp(2j\hat{\theta}) = \frac{\sum_{l=1}^{L}X_l\exp(-j\phi_l)\,Y_l\exp(+j\psi_l)}{\sum_{l=1}^{L}X_l\exp(+j\phi_l)\,Y_l\exp(-j\psi_l)}\,. \tag{12b}$$

The essential estimator for the phase angle of the transfer function involves the sum of appropriate forms of the product of the input vectors, $X_l\exp(-j\phi_l)$, and the output vectors, $Y_l\exp(+j\psi_l)$.

In a similar manner, by differentiating with respect to the gain function, H, we obtain the following least-mean-squared estimator:

$$\hat{H} = \frac{\exp(j\hat{\theta})\sum_{l=1}^{L}X_l\exp(j\phi_l)\,Y_l\exp(-j\psi_l) + \exp(-j\hat{\theta})\sum_{l=1}^{L}X_l\exp(-j\phi_l)\,Y_l\exp(+j\psi_l)}{2\sum_{l=1}^{L}X_l^2} \tag{13}$$

*Number of independent observations.
**An approximation to some value function.

Note that the estimate for the gain function, $\hat{H}$, depends on knowledge of the phase function, $\exp(j\hat{\theta})$, as well as of products of the input and output vectors.

Equation (13) can be restated as follows

$$\hat{H}\exp(j\hat{\theta}) = \frac{\exp(2j\hat{\theta}) \sum_l X_l \exp(j\phi_l)\, Y_l \exp(-j\psi_l) + \sum_l X_l \exp(-j\phi_l)\, Y_l \exp(+j\psi_l)}{2\sum_l X_l^2} \tag{14}$$

so that substitution of equation (12b) yields

$$\hat{H}\exp(j\hat{\theta}) = \frac{\sum_l X_l \exp(-j\phi_l)\, Y_l \exp(+j\psi_l)}{\sum_l X_l^2} \tag{15}$$

Equation (15) represents the least-mean-squared estimate of the system transfer function.

Because Equation (15) may be expressed as

$$\hat{H}\exp(j\hat{\theta}) = \frac{\frac{1}{L}\sum_l X_l \exp(-j\phi_l)\, Y_l \exp(+j\psi_l)}{\frac{1}{L}\sum_l X_l^2} \tag{16}$$

the numerator can be recognized as the Cross-Spectral Density estimate $\hat{S}_{xy}$ of the input and output, or

$$\hat{S}_{xy} = \frac{1}{L}\sum_{l=1}^{L} X_l \exp(-j\phi_l)\, Y_l \exp(+j\psi_l)\ , \tag{17}$$

while the denominator is the Auto-Spectral Density estimate $\hat{S}_{xx}$ of the input signal, or

$$\hat{S}_{xx} = \frac{1}{L}\sum_{l=1}^{L} X_l^2\ , \tag{18}$$

so that the transfer-function estimator becomes

$$\hat{H}\exp(j\hat{\theta}) = \frac{\hat{S}_{xy}(f)}{\hat{S}_{xx}(f)} \tag{19}$$

This is simply the ratio of the cross spectrum to the auto spectrum of the input function.

For the sake of completeness, we should mention that these vector functions may be expressed in terms of rectilinear coordinates representing in-phase and out-of-phase (co and quad) components by means of the following identities

$$\exp[\pm j\beta(f)] = \cos\beta(f) \pm j\sin\beta(f) \tag{20}$$

and $$Z(f)\exp[\pm j\beta(f)] = Z(f)\cos\beta(f) \pm jZ(f)\sin\beta(f) = Z(\pm jf) \tag{21}$$

so that $$H(jf) = \frac{\hat{S}_{xy}(f)}{\hat{S}_{xx}(f)} \tag{22}$$

and $$\hat{S}_{xy}(f) = \frac{1}{L}\sum_{l=1}^{L} X_l(jf)\, Y_l(-jf) \tag{23}$$

$$\hat{S}_{xx}(f) = \frac{1}{L}\sum_{l=1}^{L} |X_l(jf)|^2 \tag{24}$$

Figure 4 is a photograph of a transfer function of an electromechanical network (vibration exciter or shaker system) estimated in this manner on a Time/Data Model 100 Rapid Fourier Analyzer. The upper trace shows the in-phase and the lower out-of-phase or quadrature components. The horizontal axis is frequency.

Reliability of Estimate

Finally, the error in estimation can readily be determined from equation (15) by substituting equation (8), so that

$$\hat{H}\exp(j\hat{\theta}) = \frac{\sum_l X_l \exp(-j\phi_l)\,[H\exp(+j\theta)\,X_l\exp(+j\phi_l) + N_l\exp(+j\psi_l)]}{\sum X_l^2} \tag{25}$$

or

$$\hat{H}\exp(j\hat{\theta}) = H\exp(j\theta) + \frac{\sum_l X_l N_l \exp[j(\psi_l - \phi_l)]}{\sum X_l^2} \tag{26}$$

Therefore, the error in estimation, η, introduced by this procedure, is

$$\eta = \frac{\sum_l X_l N_l \exp[j(\psi_l - \phi_l)]}{\sum_l X_l^2} \tag{27}$$

This tends to approach zero as the number of observations increases indefinitely, provided that the input and noise signals are uncorrelated, because the numerator would represent the sum of a large number of vectors having random phase angles $(\psi_l - \phi_l)$, while the denominator would be a real positive definite quantity that increases linearly with the number of observations.

The mean-squared error, $\overline{\eta^2}$, in measurement can be shown to be

$$\overline{\eta^2}(f) = \frac{1}{L}\,\frac{S_{nn}(f)}{S_{xx}(f)} \tag{28}$$

where $S_{nn}(f)$ is the equivalent output-noise spectral density. The mean-squared error is inversely proportional to L, the number of independent observations entering into the estimate.

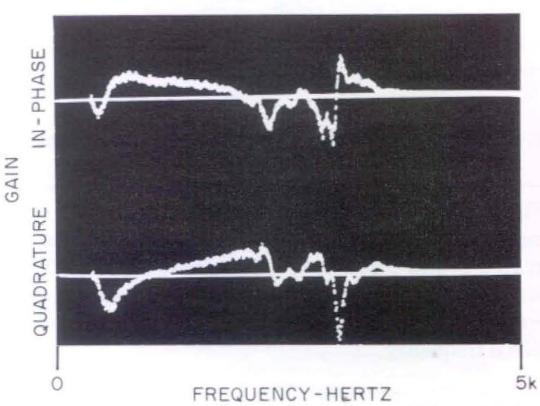


Figure 4. Transfer function of vibration-exciter system.

This error involves knowledge of the spectrum of noise; therefore, we must estimate the error by estimating the noise spectrum. The noise spectrum is approximately

$$\hat{S}_{nn}(f) = \frac{1}{L}\sum_{l=1}^{L} N_l^2(f) \tag{29}$$

and, by equations (9) and (16), becomes

$$\hat{S}_{nn}(f) = \hat{S}_{yy}(f) - \frac{|\hat{S}_{xy}(f)|^2}{\hat{S}_{xx}(f)} \tag{30}$$

$$\text{where } \hat{S}_{xy}(f) = \frac{1}{L}\sum X_l Y_l \exp[j(\psi_l - \phi_l)] \tag{31}$$

$$\hat{S}_{xx}(f) = \frac{1}{L}\sum X_l^2 \tag{32}$$

$$\hat{S}_{yy}(f) = \frac{1}{L}\sum Y_l^2 \tag{33}$$

The mean-squared error, in terms of measurable quantities, becomes

$$\overline{\eta^2} \approx \frac{1}{L}\,\frac{1}{\hat{S}_{xx}(f)}\left[\hat{S}_{yy}(f) - \frac{|\hat{S}_{xy}(f)|^2}{\hat{S}_{xx}(f)}\right], \tag{34}$$

$$\text{or } \overline{\eta^2} \approx \frac{1}{L}\,\frac{\hat{S}_{yy}(f)}{\hat{S}_{xx}(f)}\left[1 - \frac{|\hat{S}_{xy}(f)|^2}{S_{xx}(f)\,S_{yy}(f)}\right] \tag{35}$$

$$\text{or } \overline{\eta^2} \approx \frac{1}{L}\,\frac{\hat{S}_{yy}(f)}{\hat{S}_{xx}(f)}\,[1 - \Gamma(f)], \tag{36}$$

$$\text{where } \Gamma(f) = \frac{|\hat{S}_{xy}(f)|^2}{S_{xx}(f)\,S_{yy}(f)} = \frac{S_{yy}(f) - S_{nn}(f)}{S_{yy}(f)}. \tag{37}$$

From this it is apparent that

$$0 \lesssim \Gamma(f) \lesssim 1. \tag{38}$$

The function, $\Gamma(f)$, is known as the Coherence Function. It is zero when the output spectrum is due entirely to noise ($S_{yy} = S_{nn}$), and unity when the system is noise-free ($S_{nn} = 0$). This is a popular measure for the reliability of a transfer-function estimate.

In summary, we have found that the estimator that minimized the effects of spurious signal or noise sources involved measurements of the magnitude and phase of the input and output signals for each frequency of interest. These measurements could, in principle, be obtained by means of voltmeters and a phase meter. However, if the number of frequency points that must be used to describe the transfer or impedance function is large, then more efficient estimates can be made by means of digital-signal processors using fast Fourier-transform techniques. The T/D Models 1923A Time-Series Analyzer, 1923B Real-Time Fast Fourier-Transform Analyzer, or 1923C Fast Fourier-Transform Analyzer, Figure 5, directly perform the operations indicated in equation (19), utilizing the Time/Data 90 Rapid Fourier Processor. The 1923A, for example, will perform a transfer-function analysis for 128 equally spaced frequencies in less than 10 milliseconds per complete estimate.

– E. A. Sloane

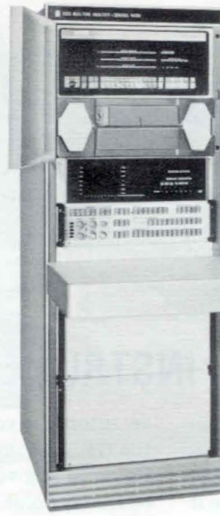


Figure 5. T/D 1923-C Real-Time Analyzer.

For Further Information

Bendat, J. S., and Piersol, A. G., *Measurement and Analysis of Random Data,* John Wiley and Sons, 1966.

Lee, Y. W., *Statistical Communications Theory,* John Wiley and Sons, 1966.

Sloane, E. A., "An Introduction to Time-Series Analysis," *Monographs I, II, and III,* Time/Data Corporation.

The Honorable Society

C. E. White

A. P. G. Peterson

On June 16, 1970, the National Conference of Standards Laboratories presented its first Awards for Outstanding Service to three members of the organization. One of the recipients was Charles E. White of General Radio. While presenting the award plaque, NCSL Chairman J. L. Hayes made the following citation:

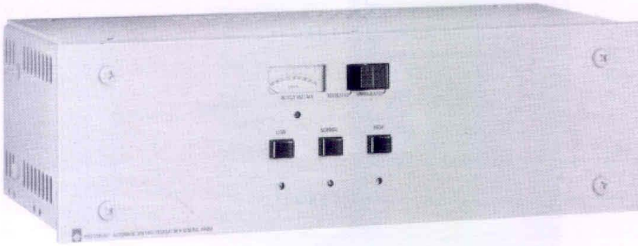
"Mr. Charles E. White has unselfishly devoted his time and energies to formulating, editing, and sustaining the operation of the *NCSL Newsletter* for the past eight and one-half years. Much of the credit for the growth of this publicity and information media is reflected upon Mr. White, who served as chairman of the NCSL Newsletter Committee and Editor of the *Newsletter.*"

The Audio Engineering Society has announced the impending award of Fellow membership to Dr. A. P. G. Peterson of General Radio. Presentation will take place at the annual Awards Banquet in New York on 14 October 1970.

Dr. Peterson is well known in the field of acoustics. He received the John H. Potts Memorial Award from the AES, in 1968, for outstanding achievement in the field of audio engineering. He is, also, a Fellow of the Institute of Electrical and Electronic Engineers and of the Acoustical Society of America.

Deviations from Accuracy

We are aware of two small errors that crept into the March-June, 1970 issue. The first, on page 5, should have translated the test pressure of 1000 microns to 1 mmHg. The second, on page 18, should have specified that switching time of the GR frequency synthesizer had been reduced to 200 μs (not 200 ms). Sorry!



NOT A CINDERELLA INSTRUMENT!

The fate of ordinary, practical, well-known, and widely-used instrumentation apparently is relegation to drudgery work and to near-obscurity when it comes to publicizing new models. The apparent lack of new-design innovation is compounded by the manner in which the general public disregards ordinary but essential instruments. We've decided to challenge this attitude by writing about the new GR 1592 Variac® automatic voltage regulators with the attitude of "How can you do without it?"

If you are a typical reader, you have control over, or access to, one or more racks of test equipment. Conceivably, each instrument is well engineered and has its own regulated power supply. Consequently, you foresee no need for a second, or master, regulator. Consider for one moment, however, that all your instruments may operate satisfactorily with an ac supply voltage of 105 volts. Modern voltage regulators usually control any voltage above the nominal 105 volts by means of a semi-conductor power-dissipating circuit to drop the voltage close to the nominal value.

In the case cited above, under normal circumstances a fair amount of heat is dissipated within the rack of instruments. Expand the number of racks and you develop the need for an air-conditioned environment to keep the mean-time-between-failure rate of the instruments from increasing rapidly. It would appear to be good logic, therefore, to reduce the supply voltage to 105 volts nominal, and to maintain continuous operation at that level. Such a step would reduce the amount of heat generated and greatly increase the useful life of the instruments. Installation of an automatic voltage regulator such as the GR 1592 maintains the instrumentation supply voltage at the desired *low* operating point.

Note that *all* voltage regulators are *not* capable of doing this. Constant-voltage transformers and reactor-type regulators operate solely to maintain the instrumentation supply voltage at the normal voltage established by the power company's distribution transformer. The GR 1592 is an electro-mechanical regulator whose output is controlled by a servo-driven Variac adjustable autotransformer of proven capability.

There is a more important point to be considered; the GR 1592 does not introduce distortion into the instrument supply voltage – a factor ignored too often by customers of regulators. In a previous *Experimenter* article,[1] we mentioned that the GR electro-mechanical regulator could track the average and peak values of supply voltage while actually detecting the rms value. This feature has significance in a number of situations. Take, for example, a capacitive-input dc power supply with a light load. Such a unit responds to peak supply voltages. A 3% distorted output from the regulator could cause as much as 3% change in the dc power-supply output, even though the regulator held to a specified 0.1% limit of deviation from a nominally rated supply voltage.

If the regulator were in control of a thermal device, which responds to the rms value of the supply voltage, distorted regulator output would affect operation of the thermal unit in a manner similar to the preceding example. When the regulator is used to control heavily-loaded capacitive-input instruments, or inductive-input power supplies, or just plain ordinary mechanical systems (all of which respond to the average values of the supply voltage), distorted output from the regulator can quickly lead to instrument operation inferior to rated performance.

The advent of digital instrumentation has created an awareness, among instrument users, of the devastating effects of spikes or sharp peaks in the supply voltage. False triggering of digital circuits is commonplace when supply voltages are used in common with distortion-producing instrumentation. Use of the GR 1592 as a buffering voltage supply unit to a block of digital instruments helps to reduce false digital outputs.

The GR 1592 can even be considered as a tool to help mitigate "brown-outs", so widely predicted by metropolitan power companies during peak-power-load seasons. If your local power company is forced to lower supply voltages drastically, use of the regulator will assure continued and satisfactory operation of instrumentation.

For readers with problems of supply for illumination devices, plating baths, or similar applications drastically affected by line-voltage variations, the GR 1592 is available for loads up to 10 kVA. Lighter load demands undoubtedly could be met by the 2-kVA model.

Complete specifications for the models listed below are available in GR Catalog U. A pamphlet describing current GR models of voltage regulators is available to readers. Address your request to:

Editor, *GR Experimenter*
300 Baker Avenue
Concord, Massachusetts 01742

[1] Chitouras, C., "Considerations In The Choice of a Line-Voltage Regulator," *GR Experimenter*, October 1967.

Design responsibility for the GR 1592 was shared by C. G. Chitouras and W. A. Montague.

Catalog Number	Description	Price
	1592 Variac® automatic voltage regulator	
1592-9700	120-V ±10% input	**$525.00**
1592-9701	120-V ±20% input	**525.00**
1592-9702	230/240-V ±5% input	**525.00**
1592-9703	230/240-V ±10% input	**525.00**
1592-9704	230/240-V ±20% input	**525.00**

Prices net FOB Concord, MA, USA.
Subject to quantity discount.

GR 1541 Multiflash Generator

NEW SHOES FOR AN OLD WORKHORSE

For a period of about 38* years, the principle of observing high-speed motion by means of electronic stroboscopes[1] has been implemented by portable GR equipment of various types. There has been wide acceptance by industry[2,3,4] of this instrumentation, capable of "stopping" fast motion without physical contact for quick analysis and of preserving the stopped-motion on film for later study.[5]

A technique, familiar to photographers, which has not received wide attention in engineering and research activities, is most useful in stopping the motion of extremely high-velocity actions without the use of high-speed cameras. This is the multi-flash or burst-flash technique by which a single sheet of film is exposed, in consecutive order, by a series of strobe flashes.

The same technique can be expanded, as test requirements become more complex, to provide for a pulse burst from a single stroboscope (Figure 1a) or a burst of individual flashes from multiple strobes (Figure 1b). It provides equivalent-shutter speeds of a microsecond to conventional ciné and high-speed motion picture cameras when the new GR 1541 Multiflash Generator is used.

The implementing system uses a stroboscope, multiflash generator, and still camera to provide frozen action on a single Polaroid** film within 15 seconds of the event. A series of complete images, each uniquely positioned on the film in time and space sequence, is available for study and action.

A Glance at the Features

Among the numerous features of the GR 1541 generator are:

- Flash bursts, adjustable in numbers and intervals
- Versatile trigger circuit, designed to accept a variety of inputs in terms of signals and connectors
- Flash intervals that can be calibrated
- Adaptability to existing stroboscopes
- Small, light, and rugged construction
- Highest intensity retained at 10-μs intervals
- Burst mode provides for initial signal to activate contact-bounce and noise-rejection circuits.

"Stopped" In Its Tracks

Many applications in motion-analysis work have been developed during the years that the stroboscope has been

*See page 18 for a partial reprint of the original GR strobe article.

**Registered trademark of the Polaroid Corp.

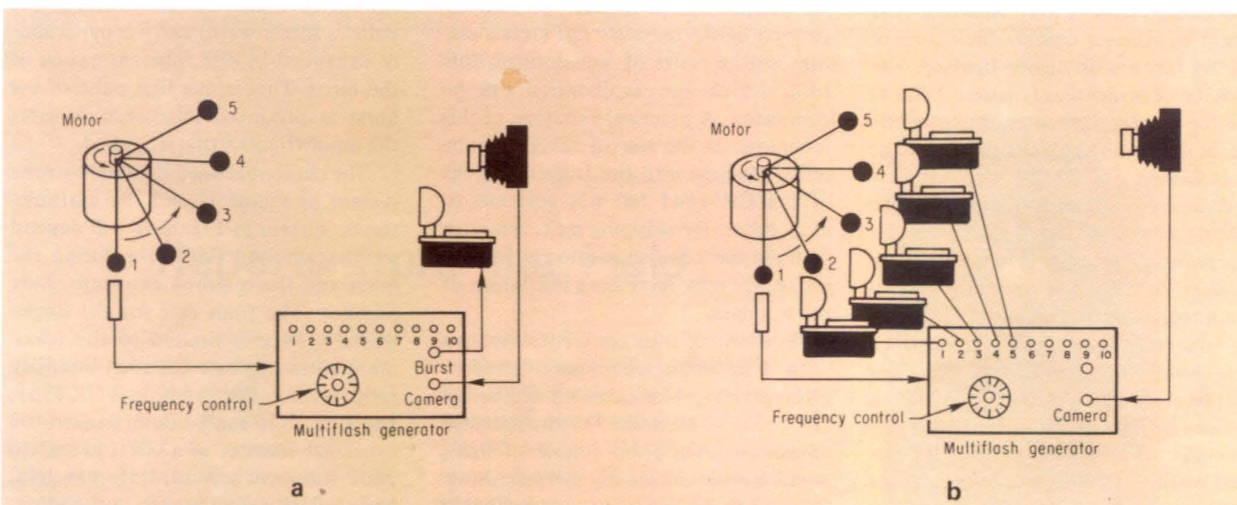


Figure 1. Flash-burst techniques. a. Single strobe unit; b. Multistrobe units.

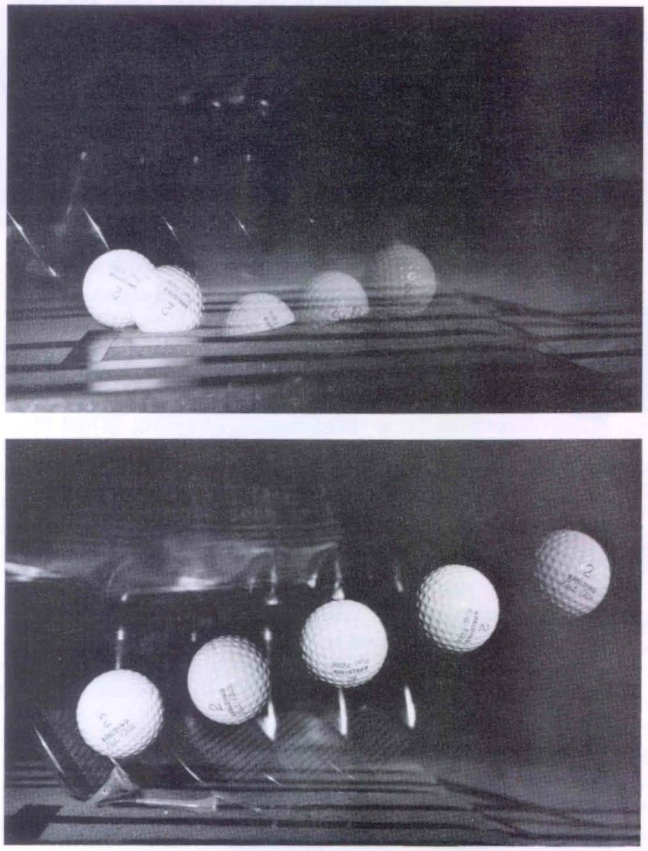


Figure 2. Golfball stroking analysis — upper, topping; lower, good lift.

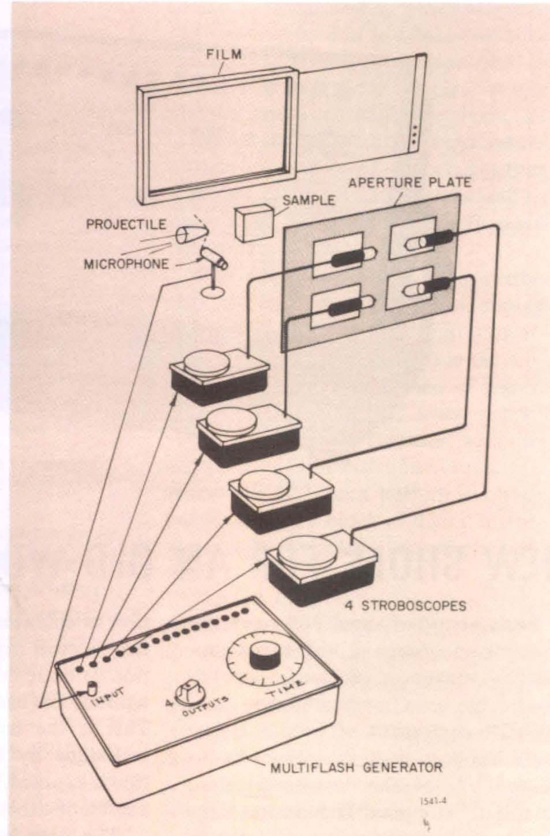


Figure 3. Instrument arrangement for projectile-motion study.

with us. Some common uses include studies of sports equipment in action, such as the impact and flight of golf balls (Figure 2) or the motion of toppling bowling pins. Athletes and would-be athletes observe their form in photos taken with strobe lighting. The velocity of projectiles (Figures 3 and 4) and the acceleration of machinery units can be determined by the use of two or more flashes.

A desirable feature in pulsed strobe lighting is that the moving object under observation get out of its own way between flashes. If it does not, successive images will overlap and "wash out" the observed action. Overlap is permissible, however, if pictures are taken for data or record purposes only, inasmuch as three or four overlapping images can normally be resolved. If color film is used, with different colored filters over each strobe light (Figure 1b), overlapped-image recognition is substantially increased.

You can observe the acceleration of a highly reflective, rotating shaft by means of another technique. The end of the shaft is painted flat black with a single peripheral white dot. Viewed by strobed light, the single dot gives a picture with a series of well-defined dots from which the acceleration can be computed. A practical variation of this technique is the use of several strobe units, plugged into the triggering jacks of the GR 1541 but not adjacent to each other. By skipping jacks, you can increase the point separation at low velocity, thereby increasing resolution of the test data.

You can separate consecutive images with a different technique – shadow photography. The compact arc in the GR 1531, 1538, and 1539 strobe lamps approximates a point source of light, which can cast unusually sharp shadows. By use of multiple, separated strobes in the system of Figure 3, the images at each succeeding interval of time fall on different areas of the film to produce a record, like that of the bullet striking the steel spring in Figure 4. Synchronization of this system is quite simple (an inexpensive microphone detects the bullet's shock wave) and it is particularly enhanced by the coherent nature of the burst. That is, the first pulse of the burst is produced microseconds after the input-trigger signal is applied.

The choice between the single-strobe system in Figure 1a and the multiple-strobe system in Figure 1b will depend on the required flash rate during the burst and the required exposure guide number. The flash rate for the single-strobe system is limited to the maximum allowable rate for each intensity range setting. For example, a GR 1531, 1539, or 1540 can be flashed up to 400 times per second, or a GR 1538 up to 2500 times per second. Unfortunately, higher flash rates require lower intensity settings; consequently, larger lens apertures are required, resulting in re-

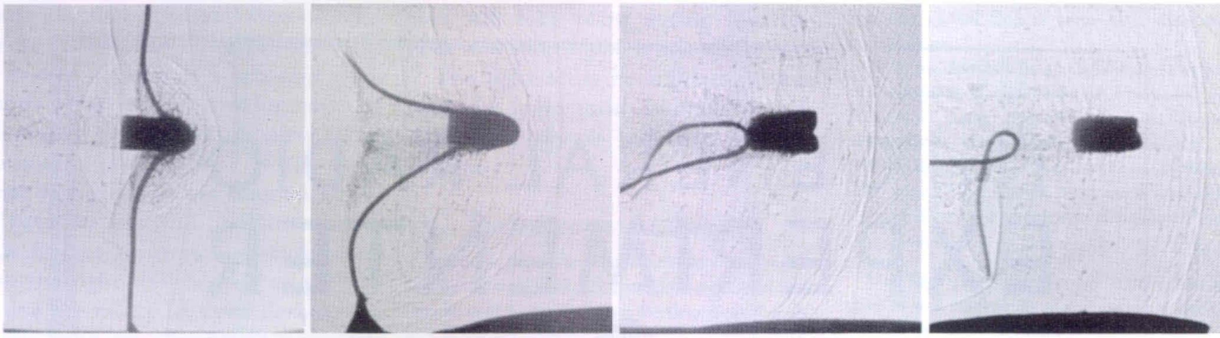


Figure 4. Multiple-flash shadowgraph sequence.

duced depth of field. This decrease in exposure corresponds to approximately 2-1/2 f-stops per intensity-range step for GR stroboscopes. When multiple strobes are used, each may be set to its highest intensity settings, thus recovering as much as 5 to 7-1/2 f-stops, with consequently increased depth of field. Each strobe is single-flashed, so that the resulting maximum burst frequency is 100,000 per second, limited only by the GR 1541 generator.

Using strobe light as the shutter eliminates blur and several distortions present even in high-speed cameras and often provides ample lighting at considerably reduced cost, weight, and line input power. For example, four GR 1538 stroboscopes and a 1541 generator can be used to produce 10,000 flashes-per-second light, about the upper speed limit for a good full-frame high-speed camera.

Some Technical Points

The GR 1541 Multiflash Generator can provide for burst groups of two to sixteen flashes, with separation between flashes continuously adjustable from 10 microseconds to 100 milliseconds. Trigger circuits in the instrument provide for a variety of input-signal sources, electrical, photoelectrical, and electromechanical. The trigger circuit is designed to reject noise or signals that occur after the initiating signal. Flash-interval control has a basic error limitation of 3%, but provision is made for a calibration signal to an external electronic counter if greater accuracy is required.

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3. Fitzmorris, M. J., "Flash-Delay Unit Simplifies Motion Analysis in High-Speed Machines," *GR Experimenter,* August 1963.

4. Holtje, M. C., "Flash! A New Strobotac® electronic stroboscope," *GR Experimenter,* April 1966.

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Development of the GR 1541 was by C. E. Miller and W. F. Rogers, who also collaborated on the above material.

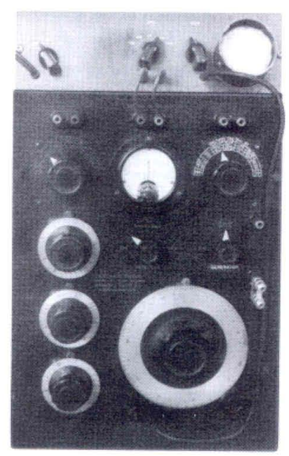
Specifications for the GR 1541 Multiflash Generator are included as a tear sheet in the back of this issue.

Catalog Number	Description	Price
1541-9701	**1541 Multiflash Generator**	**$675.00**
1541-9601	**Cable Assembly**	**7.50**

Prices net FOB Concord, MA, USA.
Subject to quantity discount.

Reports from the Field

As a nostalgic touch in the last issue of the *Experimenter*, we included a reproduction of the original article that announced the GR 650 Impedance Bridge in 1933. It brought reactions from the readers, all pleasant. One in particular, Everett Mehner of San Diego, California, was pleased to describe the manner in which he had modernized his own bridge, purchased on the surplus market three years ago. He used the battery box to house a transistorized signal generator (100, 400, 1k, 4k, and 10k Hertz) and an oscilloscope display. The latter feature permits the operator to balance null points quite simply and visually.



The GENERAL RADIO EXPERIMENTER

VOL. VII. No. 7 — DECEMBER, 1932

ELECTRICAL COMMUNICATIONS TECHNIQUE AND ITS APPLICATIONS IN ALLIED FIELDS

THE STROBOSCOPE

THE stroboscope consists fundamentally of a device which permits intermittent observations, either visual or photographic, of a moving object in such a manner as to reduce the speed of, or stop, the motion.

> THE quickness of the hand deceives the eye. But the eye knows a trick or two, and, aided by ingenious mechanisms, it is not deceived by the gyrations of machinery at far higher speeds than the trickster's hand achieves. Hence the stroboscope, which is not new, and the Edgerton* stroboscope, which is.
>
> Stroboscopes and their applications are described herewith. The Edgerton stroboscope on page 5.

The slow-motion picture is a familiar example of the interesting and profitable information which may be derived from a leisurely study of events which necessarily take place at a high rate of speed. The tennis player cannot slow the championship stroke to accommodate the laggard eye of the novice, but the camera can, and the motion picture camera is a stroboscope, but not all stroboscopes are cameras.

The camera shutter, operating at high speed, chops up the action into a number of small elements, so short that movement is not apparent in any one. The film can then be projected at normal speed with results that are instructive, or even backward with results that may be amusing. The function of the shutter is to exclude light from the film except for brief flashes. It seems reasonable that the same result can be obtained by shutting off the light from the object, except for brief flashes. This is the nature of the second style of stroboscope, of which the Edgerton type is the outstanding example. Obviously this type of stroboscope is well adapted for visual observations. Photography must still be used, if a non-repeated event is viewed, to store the elementary views and to release them later at a rate that the eye and mind can follow.

Consider, however, an indefatigable tennis player who repeats his stroke,

*The Edgerton Stroboscope is a development of Prof. Harold E. Edgerton, Massachusetts Institute of Technology.

identically, one thousand times a minute in a darkened room. If the light be flashed on him at a constant rate, exactly equal to his stroking rate, he will appear as though motionless under continuous illumination. If the flash speed be slightly slower than his stroking rate, his arm will be illuminated a little farther along in the stroke each time the light flashes and, as the eye retains the image between flashes, the madly stroking player will seem leisurely, and a single stroke can be spread over a minute if desired.

Humans, tennis playing or otherwise, cannot repeat uniform cycles at any such speed. Machines can, and wherever complicated machines are designed, built, or used, the ability to watch their operation in slow motion without photography is a boon.

The stroboscope permits stopping the motion of the machine (visually) for examination of machine or product at any part of its operating cycle while the grommets flow into the hoppers at undiminished speed. Or, perhaps, a squeaking clutch, a vibrating shaft, or a chattering valve spring stands between a new model and a waiting public—which will not wait long. A slow motion study will show the trouble, or the primary motion may be stopped and the vibrating member made as conspicuous as a mosquito-brushing hand at formal guard mount.

Sometimes the transient movement or vibration takes place at too high a speed for the eye even with the primary motion stopped. Here photography is resorted to for a second slowing down of the transient.

A little consideration of what is being done by the stroboscope is sufficient to set up the requirements of a satisfactory one.

An accurate means of timing the flash and a prompt and accurate response to the flash control are essential, otherwise the object will be viewed at irregular intervals, and vibrations not present in the object viewed will be introduced.

The flash must be of extremely short duration. Otherwise appreciable motion will take place during illumination, and blurring of detail will result.

The light must be brilliant. Otherwise the room must be made entirely dark, and details will not be seen clearly.

Stroboscope Arithmetic

Suppose that the object to be observed is executing uniformly R complete cycles of motion in unit time. Suppose further that the object is either viewed through a shutter opening for F brief, uniformly timed intervals, or is illuminated by F uniform instantaneous flashes of light in unit time. Then, if

$$R = nF \qquad (1)$$

where n is an integral number, it will be evident that each point of the object will be in exactly the same position in its cycle of motion at each observation, resulting in what we shall designate as a condition of "perfect" synchronism. Accordingly, all apparent motion of the body will be arrested, so that it will appear to be stationary at some particular phase in its cycle of motion, provided that the opening of the shutter or the flash of the lamp is of extremely short duration. If this interval of observation is of sufficient duration, the moving object, even when viewed stroboscopically, will appear blurred in outline, since each point of the body executes a perceptible amount of motion during the interval of observation.

It is further evident that the phase of the observed position of the object in its cycle of motion may be controlled at will merely by shifting the phase of the synchronous shutter or light flash with respect to the motion.

The special case of perfect synchronism, in which the frequency of motion and of observation are identical, is known as "fundamental" synchronism.

If n is greater than 1, the object will be observed only at every nth cycle of motion, so that the integrated illumination is reduced to the fractional amount $1/n$ times the illumination at fundamental synchronism.

Although any condition of perfect synchronism will completely arrest the motion, it is obviously desirable to work at the condition of fundamental synchronism.

If, on the other hand,

$$F = kR \qquad (2)$$

where k is any integral number greater than 1, then each point of the object will be visible k times per cycle of motion and will, accordingly, be observed successively at k points equally spaced, in time, throughout the cycle of motion. Such a condition, which is known as "partial" synchronism, while apparently arresting the motion of the object, is not, in general, satisfactory for visual stroboscopic observations. For example, a rotating disc having one radial line is seen as a disc with k radial lines.

A more distinct image is obtained at partial synchronism if the body is composed of mk identical parts equally spaced, in time, throughout the cycle of motion, *e.g.*, by a wheel having $P = mk$ spokes. Further, it can readily be shown that such a wheel will appear as a stationary wheel having P spokes whenever

$$PR = nF \qquad (3)$$

On the other hand, the wheel having P spokes will appear as a stationary wheel having nP spokes whenever

$$nPR = F \qquad (4)$$

Reference to equation (3) shows that there are, theoretically, an infinite number of values of R or of F for which a wheel of P spokes will be seen as a stationary wheel of P spokes. The larger the value of P, the greater will be the number of these partial synchronisms which occur within a given range of values of R or F. These facts are of importance in using the stroboscope to determine the frequency or speed of cyclic motions.

We have so far analyzed the fundamental laws of the stroboscope for conditions of exact synchronism, either partial or perfect. Consider now the case where the cyclic frequency of motion is slightly greater than an integral multiple of the frequency of observation—

$$R = nF + S \qquad (5)$$

where S is small compared to R. This means that the moving object will execute slightly more than n cycles of motion during the interval between observations so that the phase at which it is seen stroboscopically will continually advance. The object will therefore appear to move at a slow cyclic frequency of

$$S = R - nF \qquad (5a)$$

cycles in unit time and to travel in the same direction as the object is actually moving.

Conversely, if the cyclic frequency is slightly less than an integral multiple of the frequency of observation the phase at which the object is seen stroboscopically will continually recede so that the object will appear to move at a slow cyclic frequency in a direction opposite to the true motion:

$$S = nF - R \qquad (6)$$

The slow stroboscopic motion which can be obtained in this manner, and which can be adjusted to become a very small fraction of the true speed, makes the stroboscope extremely valuable in watching the cycle of motion of machinery running at speeds too high to be followed with the unaided eye.

The frequency of stroboscopic motion, S, may be made as slow as desired. On the other hand if S is increased above a certain limit the observed motion becomes intermittent and less satisfactory for purposes of visual study.

— **Horatio W. Lamson**

another new product

PROGRAMMABLE DECADE RESISTOR

The GR 1435 Programmable Decade Resistor was designed for maximum customer-use flexibility consistent with accuracy and cost. The basic instrument covers the five-decade span from 10-Ω to 100-kΩ per step, with each decade a plug-in board. Mechanical and electrical provision has been made to allow simple conversion to a six- or seven-decade instrument, should the need arise. Reed switches used throughout the instrument are of the miniature mercury-wetted type, for low and repeatable zero resistance as well as bounce-free operation. The high and low terminals of the resistors are isolated from ground; this permits use where a floating resistor is required.

Three distinct modes of operation are provided for the user's convenience:

Manual Mode The desired resistance is set on the front-panel dials, just as one would set a conventional decade resistor. This is useful, for example, when you are making accuracy checks on the GR 1435, to determine how many decades need be remotely controlled for a particular application, or when you are manually checking proper system operation during initial set-up stages.

Manual/Remote Mode Some of the decades may be set to the "R" position on the front-panel dials and be remotely controlled, while the remaining dials are set to a particular value of resistance and held constant. This has the advantage of requiring four less control lines for every decade which is manually set.

Remote Mode You can select this mode by turning all decade dials to the "R" position; by turning the power switch to the "REMOTE" position; or by applying logic "0" to pin V of the rear-panel connector. Resistance is set by application of negative true 1-2-4-8 BCD signals at standard DTL or TTL levels, or contact closures to ground, to each decade via the 36-pin rear-panel connector.

A feature that deserves special mention is the ability to short or open the decade resistor terminals remotely. Grounding pin 18 of the 36-pin rear-panel connector shorts the resistor terminals, while grounding pin 17 opens the resistor terminals. This is particularly useful if discontinuities are objectionable, which may exist when resistance settings are changed (the worst case being the change between "7" and "8" where all four reeds change state). With proper timing, the resistor terminals could be either opened or shorted during this switching interval, whichever is more pertinent to the application.

A few typical applications are illustrated in the figures, ranging from a simple programmable amplifier load to a programmable oscillator or time constant.

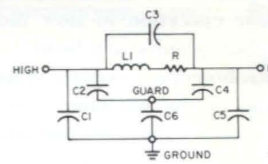
Catalog Number	Description	Price
	1435 Programmable Decade Resistor	
1435-9700	Bench Model	**$750.00**
1435-9701	Rack Model	**730.00**

Prices net FOB Concord, MA, USA.
Subject to quantity discount.

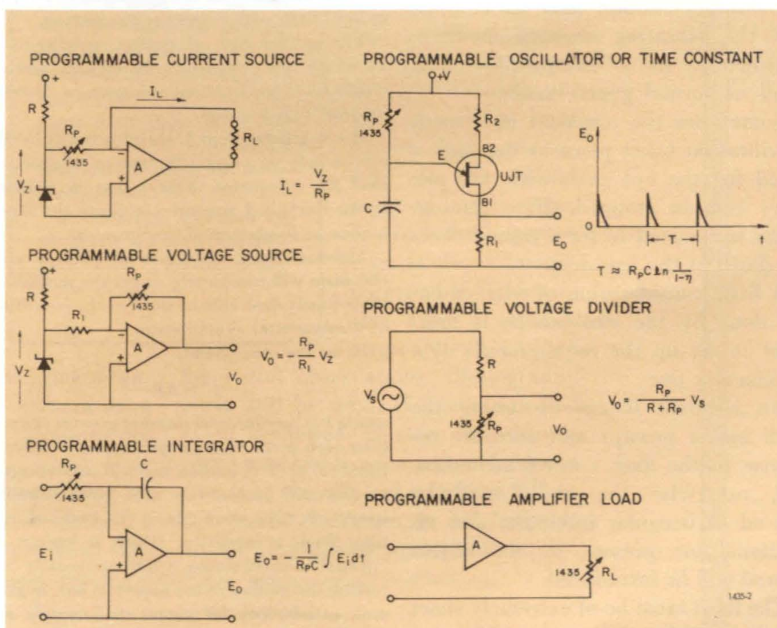
This instrument was designed by Peter Gray of GR's Component and Network Testing Group, who contributed the material for this article.

Complete specifications for the GR 1435 are in GR Catalog U; minor revisions are shown below.

Frequency Characteristics: At high-resistance values, frequency characteristics depend mainly on capacitances and on the type of connections used (2- or 3-terminal, grounded or guarded). At low resistance values, they depend mainly on the inductance. Calculations based on values shown should give approximate series-resistance error.



Parameter	Decade Resistance R = 100 kΩ	R = 1 MΩ
C1	19 pF	11 pF
C2	76 pF	23 pF
C3	19 pF	16 pF
C4	247 pF	276 pF
C5	46 pF	51 pF
C6	1606 pF	1606 pF
L1	23 μH	23 μH



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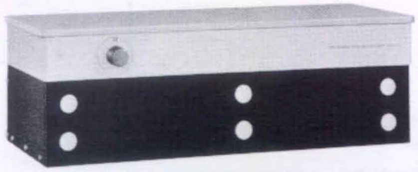
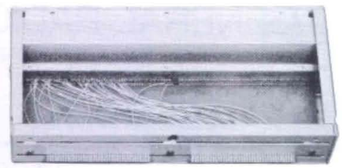


Universal Device Adaptor.



Underside of 1790-9603 Standard Device Adaptor Kit showing socket holes and typical wire-wrapped connections.

GR 1790–MORE VERSATILITY AND CAPACITY

Versatility is the name for the universal device adaptor recently made available to GR 1790 customers.

The logic-circuit analyzer has proven itself facile enough to handle most logic-circuit test problems presented it simply by means of suitably adapted interfacing.

GR engineers incorporated several important principles in the design of the interface adaptors. Connections within the adaptor are made simply, by wire wrapping to terminals mounted on printed boards. Auxiliary control or monitoring circuitry and loads are easily connected within the adaptors.

Customers are given the choice of a standard adaptor or of a universal adaptor. In the universal adaptors, input and output connections can be determined by the test program; this permits acceptance of a greater variety of devices as well as providing checks of outputs and inputs. Use of the universal adaptor permits tests for shorted inputs.

The adaptors are easily inserted into or removed from the analyzer by action of a single lever. Provision is made on one standard board to mount sockets in rows 0.250 inch apart and spaced at 0.125-inch internals. Adaptors are available with 24, 48, 72, and 96 inputs and 48, 72, 96, 120, and 144 outputs in predetermined combinations.

Complete specification details for the GR 1790 adaptors are available on the tear sheet at the back of this issue.

As customers for the standard GR 1790 Logic-Circuit Analyzer become familiar with its operation, we anticipate their needs will grow to expand its application to more complex tests. Or, a need for expanded memory storage will be evident as test programs lengthen. It is even possible that originally limited funding for capital expenditures may be increased as the savings, made possible by the GR 1790 in action, are brought to management's attention. For any of these reasons, GR is prepared to help its customers expand their standard analyzers by supplying and installing several retrofit kits.

Kits are available in two basic formats – one to expand memory by 50 times (Option 2) and the other to provide capability for the addition of programmable logic levels and programmable power supplies (Option 3). Both options are available in 50-Hz as well as 60-Hz versions, to accommodate overseas customers. Options are also available separately or combined. Neither option requires more physical space.

The kits will be installed by GR district office service-department personnel. Training required for operation with Option 2 will be provided the customer; no further training is required for Option 3.

Complete specification details for the GR 1790 options are available on the tear sheet at the back of this issue.

Description	Price
1790 Logic-Circuit Analyzer	
Retrofit Kits, for installation of options in the field by GR personnel	
Kit 2AK for Additional Memory, 60-Hz line	**$10,500.00**
Kit 2BK for Additional Memory, 50-Hz line	**10,600.00**
Kit 3K for Programmable Levels, 50 to 60-Hz line	**13,500.00**
Kit 2A-3K for both options, 60-Hz line	**23,000.00**
Kit 2B-3K for both options, 50-Hz line	**23,100.00**

	Dedicated Inputs	Dedicated Outputs	Programmable Inputs/Outputs	Price
Standard Device Adaptor Kits				
1790-9601 no socket holes	72	72	—	**$ 130.00**
1790-9602 no socket holes	96	144	—	**195.00**
1790-9603 socket holes	72	72	—	**135.00**
1790-9604 socket holes	96	144	—	**195.00**
Universal Device Adaptors				
1790-9605 no socket holes	72	120	24	**600.00**
1790-9606 no socket holes	48	96	48	**825.00**
1790-9607 no socket holes	24	72	72	**1000.00**
1790-9608 no socket holes	—	48	96	**1350.00**

Prices net FOB Concord, MA, USA.
Subject to quantity discount.

Recent Technical Articles by GR Personnel

"The Human Factor in Precise Measurements," C. E. White, *Measurements and Data,* March-April 1970.*

"Semi-Automatic DC-DVM Calibrator," R. P. Anderson, *Measurements & Data,* May-June 1970.**

"Computer Aids Redundant Logic Search," G. R. Partridge, *Electronic Design News (EDN),* 15 June 1970.**

"Planning Investments in Research and Development," W. D. Hill, *Managerial Planning,* July/August 1970.*

"A Noise Exposure Meter," G. R. Partridge, to be presented 3-6 November, Acoustical Society of America.**

"Approximate Transfer Characteristics of a Condenser Microphone with Diaphragm Stretched Over Raised Points of the Backplate," S. V. Djuric, to be presented 3-6 November, Acoustical Society of America.**

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Our Cover is an illustration of industry in action - with a G-S 1703 Audiometer. This picture is being repeated in many plants these days. The subject is undergoing a hearing test to detect any deviation from the normal response.

Inevitably, as time goes by, we awaken to the fact that our hearing is not as good as it used to be. If we are lucky, Nature will proceed **slowly**, but surely, to steal this sense from us. If we are unlucky, Nature will receive much help in ***hastening*** the process from the trappings of civilization - the noises of machinery, jet engines, rock and roll, etc. The tragedy of the latter case lies in the insidiousness by which we are consigned to that quieter world.

Deafness was recognized in the practice of medicine from its inception but comparatively little was known about how to measure the loss until the early part of this century. In line with Lord Kelvin's axiom concerning measurement and knowledge, the medical profession has contributed much data since 1900 to establish degree and type of deafness in patients. These data were derived from the **audiometer**, an instrument which has become increasingly familiar as industry has been alerted to one of its responsibilities – the well-being of the worker.

In this issue, Rufus Grason, president of GR's subsidiary, Grason-Stadler, describes the foundations of audiometry and the development of the audiometer.

Audiometers can be complex or simple, as established by their applications in hearing research or for clinical studies. But they are indispensable in this era of an alerted public and a benevolent judiciary, determined to prevent man from helping Nature's relentless but slow progression toward an awesome, silent world.

C E White

C. E. White
Editor

Another year draws to a close. Perhaps it has not been the best of years to many of us but it **is** a part of our lives. Looking forward optimistically, we at GR hope to share with our readers a Holiday Season filled with happiness and a New Year of growth and prosperity.

Season's Greetings!

Audiometric Measurement: 150 Years of Applied Research

One of the many considerations influencing the recent merger of Grason-Stadler with General Radio was the former's expertise in life-science instrumentation. Of particular interest was G-S's leadership in the design and distribution of acoustic and audiometric instrumentation—an area which significantly complements that occupied by GR's acoustic measurement devices. In the past year, Grason-Stadler has introduced two new major models, the G-S 1701 Diagnostic Audiometer and, more recently, the G-S 1703 Recording Audiometer. Adding to an already extensive line, these models represent two extremes of a continuum of devices whose applications range from relatively simple automatic screening to highly sophisticated research. This article describes these two units, how they came to be and their significance in the audiometric field.

WHAT IS AN AUDIOMETER?

In simple terms, an *audiometer* is an electronic instrument used to measure an individual's hearing acuity. The simplest units perform this function by providing to the listener (usually through earphones) an audio signal (commonly a pure tone) of known intensity and frequency. More sophisticated instruments offer the listener a variety of signals, pure tones, white noise, and speech, through a variety of output transducers – earphones, bone vibrators, or loudspeakers. These audiometers often will record, for several frequencies, the intensity level at which the listener just hears the signal.

The audiometer is usually operated under the auspices of an *audiologist*, a professionally trained individual interested in the measurement of the hearing function, its relationship to normative data, its assessment and, where appropriate, its treatment. As a formal discipline, the field of audiology plus the instrumentation that accompanies it is slightly more than 20 years old. Both were precipitated during the post World War II era when thousands of service personnel with varying degrees of hearing impairments returned to civilian status. Then, more than ever before, there was need for equipment, trained personnel, and accepted, proven techniques.

G-S 1701 Audiometer

ROOTS OF AUDIOMETRY

The earliest attempts, in the beginning of the 19^{th} century to establish techniques for the measurement of hearing involved little or no instrumentation as such. On the simplest level were live-voice tests, which typically required the tester to maintain a fixed-intensity speech level while varying the test distance until he could just be heard by the listener. Although this type of test yields some information regarding how an individual handles speech communication, it requires well-practiced testers, is subject to great test-retest variability, and is, at best, adequate only for screening purposes.

Other early hearing tests involved the tuning fork, whose prongs vibrate when struck lightly, producing a pure fixed-frequency tone. The tuning fork provided a convenient means of generating a precisely repeatable fixed frequency, even though it was a greater problem to control the sound level at the subject's ear with the tuning fork than with live speech. Moreover, it also manifested the ability to transmit its signal by means other than air conduction, for, if its base is placed in contact with any solid material - wood, metal, or the human skull - it induces sympathetic vibrations in that material. This characteristic made the tuning fork uniquely suitable for early attempts to determine the anatomical site responsible for a given hearing loss.

Normally, the vibrating fork would be held outside the ear, air serving as the initial conductive medium. If the tone generated were heard by the listener, it was only because the ear was "normal" and the tone had been transmitted successfully through the entire auditory chain, including the outer, middle, and inner ears. If this test were unsuccessful, the next step might be to place the base of the vibrating fork behind the ear on the subject's mastoid bone, which serves as the initial conductive medium. For the tone generated by this process to be heard, it would only have to excite, by direct conductive vibrations, the inner-ear neural mechanisms through which acoustic stimuli are transmitted to the higher centers in the brain. If this step successfully elicited a response where air conduction had failed, it would seem to indicate some blockage or discontinuity in the outer or middle ear. This mode of determining any differential sensitivity to air-conduction and bone-conduction tests proved to be a viable diagnostic procedure. The basic technique pioneered with the tuning fork was subsequently refined and adopted as a standard procedure in the growing diagnostic repertoire.

EARLY INSTRUMENTS

For many years the tuning fork and live-voice test served as the most sophisticated means to measure human hearing. By the end of the 19th century, however, technology had advanced to a point where corollaries to these types of tests could be implemented by electro-mechanical instruments. The earliest instruments designed to test hearing were scarcely one step removed from the tuning fork, in some cases containing that very device as their central component. In at least one instance, the tuning fork's oscillations were used to modulate an electrical circuit and to produce an alternating current in a secondary circuit. Part of this secondary circuit was a telephone receiver that reproduced the frequency of the vibrating fork at the listener's ear. This technique established a signal source with repeatable frequency characteristics. Popularly called an "Acoumeter," this tuning fork audiometer served as the basic auditory-test instrument until the alternating-current generator made possible the production of a signal with a wider frequency range than that of a tuning fork. The availability of the vacuum tube, in the early 1920's, made electronic audiometers commercially feasible.

By the early years of the 20th century, then, an electro-mechanical replacement had been found for the purely mechanical tuning fork in air-conduction threshold tests. It was only a few years later that an electro-mechanical successor was found for the tuning fork in its second application – tests of bone-conduction hearing. To reproduce the effect of the vibrating base of the tuning fork, the diaphragm of a telephone receiver was replaced by a strip of metal to whose surface was attached a metal rod. This device, driven by the same auditory signal used in the air-conduction tests, could now serve as the vibration source.

The electronic successor to the earlier live-voice speech testing came about gradually through the early years of the 20th century. By 1927, the technique of recorded speech tests – implemented by a spring-wound phonograph – had reached a new peak of sophistication in the Western Electric 4A Audiometer. This unit permitted individual subjects, or even entire classes of subjects, to be given speech threshold tests.

Few really significant audiometric developments took place in the 30's and early 40's, prior to the outbreak of WW II. The field was growing, however, and commercial audiometers appeared in increasing numbers on the market, although their main characteristics were really quite similar. They were, without exception, vacuum-tube based. Several were equipped with sweep-frequency oscillators, though the majority provided only a limited number of frequencies, most of them the so-called "tuning-fork frequencies" of 128 Hz and its multiples. Output transducers generally included a variety of types of earphones and early renditions of the bone vibrator. An electric buzzer was included as a rough approximation of a masking source on many units, to "mask" or shield the ear not under test from signals conveyed by air or bone from the ear under test. Intensity was usually specified in terms of decibels of attenuation for each frequency used.

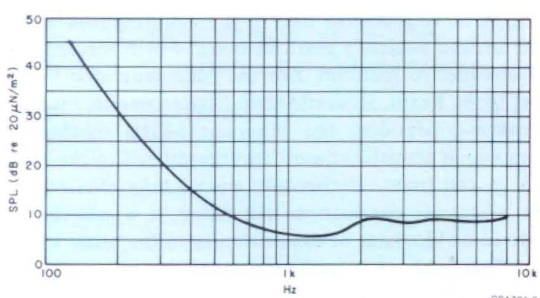

Figure 1. 1964 ISO threshold values for pure tone. Earphone reference based on measurements made on National Bureau of Standards 9-A coupler and Western Electric 705 earphone.

INSTRUMENTATION IMPROVEMENTS

In terms of the test equipment that had preceded them, the instruments described above reflected significant advancements in both audiometry and general electronic technology. The technology as a whole, however, was still in relative infancy.

Standards

When audiometers first were commercially manufactured, there were no accepted standards to specify either "normal" thresholds, acceptable signal parameters, or test techniques. Over the years, however, many organizations have been formed specifically to establish, revise, and maintain such standards.

One of the most important standards, worked out over the course of several years, specifies the "normal" threshold intensities of the most significant frequencies in the audible continuum. These so-called normal absolute threshold values were obtained by screening large segments of the population to locate individuals without obvious hearing abnormalities, then by meticulous tests of these individuals' hearing. After further screening of the data, a statistical average was made and an absolute threshold value for each of several frequencies determined. Figure 1 shows the ISO pure-tone absolute threshold levels versus frequency.

On modern audiogram forms, such as that from the G-S 1701 (shown in Figure 2), an individual's hearing at several frequencies is plotted with reference to Hearing Threshold Level; 0 dB HTL is equivalent to the standard normal threshold values shown in Figure 1.

The G-S 1701, like most other audiometers, changes the intensity of the output as the frequency is changed and automatically references the signal to the accepted threshold standards. A subject with hearing acuity more sensitive than normal will show a negative HTL. A subject whose hearing is less sensitive than normal will show a positive HTL, i.e., he will require a signal more intense than normal to hear the same frequency tone.

Calibration

The American National Standards Institute, Inc. (ANSI) the National Bureau of Standards, and other regulatory groups also have tried to establish standards in the critical area of audiometer calibration. The output level of early audiometers, for example, was calibrated by the measurement of voltage produced across the earphones. While such an approach could accurately describe the adequacy of the signal

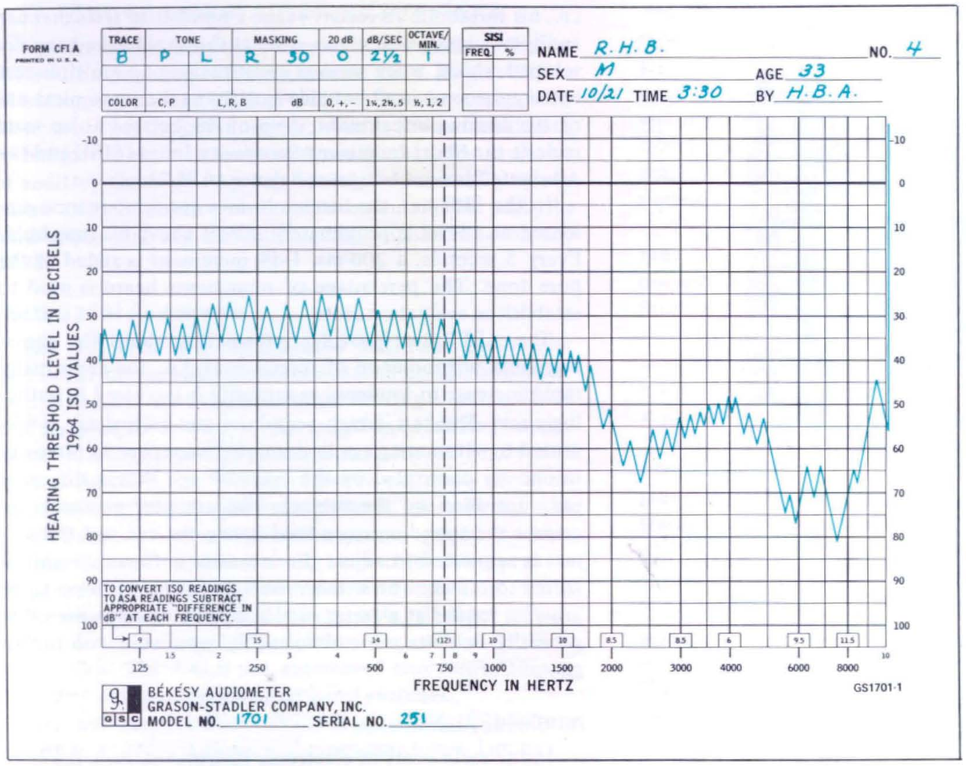

Figure 2. G-S 1701 audiogram, illustrating possible inner-ear hearing loss.

within the system, it took into account neither the likely non-uniformity of the earphone response nor the effect on that response of the volume, resonance, and impedance characteristics of the ear into which the signal was directed.

The most significant improvement in this area came when couplers and, later, artificial ears were introduced into the audiometric calibration process. Both these devices, made with known volume and material, serve as substitutes for the human ear. The earphone of the audiometer to be calibrated is tightly fitted to the mouth of the coupler or artificial ear. The sound-pressure level of the earphone signal, introduced at a fixed input level, can then be measured and read through a microphone contained in the cavity.

Masking Sources

The importance of a masking source to mask, or block, transmission of the test signal to the ear not under test was recognized in the 1920's, and early audiometers contained an ordinary electrical buzzer specifically for this purpose. It soon became apparent that the precise nature of the masking agent – especially its frequency spectrum – significantly affected the pure-tone threshold being measured.

White noise, whose spectrum contains equal amounts of all audible frequencies, provides a more effective masking agent than the buzzer and continues to be used to the present day. White noise masks all frequencies equally, including that of the test signal, and with a minimum production of beats or harmonics. This major advantage of white noise, however, is also its main disadvantage. Because the white-noise spectrum is so broad, it introduces to the ear not under test a much higher over-all energy level than is required to mask any given pure tone. Ideally, most efficient masking would be accomplished with a very narrow band of frequencies centered around the test tone, which would concentrate the available energy in the vicinity of that tone.

As its latest approximation to an optimum masking signal, Grason-Stadler has incorporated into its 1701 Audiometer a variable narrow-band noise source whose bandwidth changes as a function of the frequency of the test tone. This variable bandwidth, which permits efficient masking to take place at all frequencies, is unique to the G-S 1701 and stands out as one of its most important features.

Speech Testing

The disadvantages of live-voice tests have already been mentioned. It became apparent at a relatively early stage, however, that whatever the advances in sophistication of pure-tone audiometry, speech material could not be abandoned entirely as an audiometric stimulus. Not only are there psychological advantages in tests with a speech stimulus, normally dealt with by average listeners, but there are certain common hearing disorders in which the subject manifests significantly less ability to understand speech than his pure-tone threshold would suggest. For these reasons, improved instrumentation and the standardization of testing materials were needed.

A variety of speech materials has been developed specifically for speech audiometry through the years. Much of the work originated at the Harvard Psychoacoustic Laboratory, the Bell Telephone Laboratories, and the Central Institute for the Deaf. Such material generally includes the equally stressed (spondee) and phonetically balanced lists whose

Rufus L. Grason received a very practical introduction to the field of psychoacoustics and its instrumentation by first working in, and later assuming responsibility for, the electronics shop at the Harvard Psychoacoustic Laboratory. His experience and training lead him in 1949 to form, with Steve Stadler, the Grason-Stadler Company, now a subsidiary of General Radio. As President and Director of Engineering, he has been intimately involved with design and development of the company's equipment, most recently the 1701 and 1703 audiometers. In the early 1960's, Mr. Grason served as a member of the American Standards Association writing group whose recommendations for audiometers resulted in the ANSI 1969 specifications. He is currently Secretary of a subcommittee of the International Electrotechnical Commission, preparing specifications for diagnostic and research audiometers.

phonemic make-up roughly matches that of American colloquial speech. This material is presented in standard speech tests either by an operator or through tape or phonograph inputs.

Modern audiometers such as the G-S 1701 include a VU meter that can be switched into the circuit to calibrate the input signal, either the speaker's voice or a recorded (tape or phonograph) input. To deliver a fixed-intensity speech signal to the subject in live-voice tests, the operator need only speak into the microphone and control the intensity of his speech with the aid of the monitor VU meter. Precisely repeatable intensity increments or decrements can then be made by a simple adjustment of the attenuator control.

A second method of implementing speech tests is to present the recorded material by means of either a phonograph or a tape recorder. To facilitate such an approach, most of the standard word lists are now provided by audiometer manufacturers on records, which have the advantage of providing a uniform speaking voice; this permits excellent inter-clinic data comparisons. At the beginning of these records a 1000-Hz calibration tone is generally included, which can be used to ensure that different testers present subsequent test materials under more nearly comparable conditions.

Suprathreshold Tests

One of the major disadvantages of the earlier audiometers was their exclusion of suprathreshold testing. Almost without exception, audiometers were used to determine the minimum audible intensity that could be detected by the listener, i.e., his threshold. In recent years, a number of tests that use auditory stimuli well above normal threshold have been developed which, when used as constituents of a multiple-test battery, become useful as aids in defining the anatomical site of the hearing impairment. Two of the better-known tests include the Short Increment Sensitivity Index (SISI) and the Alternate Binaural Loudness Balance (ABLB).

In the SISI test, the listener hears a continuous tone presented at a level approximately 20 dB above his threshold. Every 5 seconds, a 200-ms, 1-dB increment is added to the pure tone. The percentage of increments heard is used to establish an evaluation score.

The ABLB test provides information about the suprathreshold phenomenon of recruitment, i.e., the abnormally rapid increase in loudness as intensity is increased in pathologic ears. This test, which postulates one normal ear, is presented by alternating a pulsed tone between ears, its intensity in one ear controlled by the operator and that in the other ear controlled by the subject. The operator gradually increases the sound-pressure level in the one ear, and the subject is requested to adjust the intensity in the other until it seems to match. The listener who perceives the tone to be growing louder at a faster rate in one ear than in the other generally exhibits some abnormality associated with the organ of Corti.

Automatic Audiometers

In the early years of electronic instrumentation, the audiometer was manually operated. Intensity and frequency changes, and any timing of signal duration, were implemented by the operator. The responses of the subject, usually a verbal "yes" or "no" or a hand signal, were also manually recorded by the operator.

In recent years, especially since WW II, an increasing number of these functions have been automated. In addition to the obvious benefit of operational ease, such automation has resulted in increased reliability by presenting standard test sequences, free from operator intervention and the consequent possibility of error. On the automatic G-S 1701, for example, standard tests that employ short auditory signal presentations are automatically timed. The G-S 1701 also varies intensity and frequency parameters, and it records the subject responses to these stimuli. The technique through which this is accomplished is generally referred to as the Békésy technique, after Georg von Békésy who developed the procedure in the 1940's.

Békésy's technique requires that the subject's threshold be recorded continuously at several test frequencies. While the test is being administered, a recording pen is moved along the horizontal axis of the audiogram form, on which frequency is plotted. In the absence of a subject response, an automatic attenuator associated with the subject switch increases the sound level and simultaneously moves the recording pen down along the HTL (vertical) axis of the form. In the presence of a subject response, the attenuator decreases the sound level and moves the recording pen up along the chart's vertical axis. The end result of this procedure is that a record is made of the subject responses in the region between audibility and inaudibility.

THE GRASON-STADLER AUDIOMETERS

Grason-Stadler has been designing and manufacturing audiometric equipment for more than 20 years. Its present line of audiometry-related instrumentation includes a speech audiometer; a psychogalvanometer, which utilizes conditioning techniques to elicit a change in skin resistance as an indicator of auditory threshold; a group hearing aid, essentially an amplifier used in group situations to communicate with the hard-of-hearing; and two audiometers – the G-S 1701 and 1703.

The G-S 1701 Audiometer

The G-S 1701 is a sophisticated diagnostic audiometer used under the auspices of professional audiologists to measure and assist in the evaluation of the hearing function. Since it is designed to be used in the diagnosis of hearing impairments – which requires the administration of whole batteries of related tests – the instrument is extremely versatile. Signal sources include pure-tone, white, narrow-band and speech noise, as well as microphone, phonograph, and tape recorder inputs. Output transducers include loudspeakers for sound-field tests, earphones for air conduction, and a bone vibrator for bone-conduction tests, all three advantageous for reasons mentioned above. Suprathreshold tests such as SISI and ABLB are automated and can be implemented by changes of a few front-panel switches.

Intensity output of both channels of the G-S 1701 is from −15 dB to +115 dB HTL for mid-range pure tones. Timing for other tests can be implemented manually or automatically in a variety of switch-selected modes. To facilitate verbal communication with the listener being tested, a talk-forward/talk-back system with independent level controls is included with the G-S 1701.

Perhaps its most outstanding feature is the flexibility of its automatic control provisions. In addition to sweep-frequency Békésy, it can present fixed-frequency Békésy or automatic ABLB tests. Another distinctive feature is its variable bandwidth masking source, the first such masking source to appear on a commercially available audiometer.

The G-S 1703 Audiometer

The G-S 1703 Recording Audiometer is much simpler and less sophisticated than the 1701. It has been designed primarily for use in the early stages of a well-developed hearing program, to distinguish normal from hard-of-hearing individuals. Although the G-S 1703 will be used for a variety of applications, it will undoubtedly find wide-spread use in industrial and business situations where high ambient-noise levels might adversely affect hearing.

The existence of noise-induced hearing loss has been recognized for a number of years; the first national conference on noise was held in 1952. Since then, numerous variables contributing to noise-induced hearing loss have been determined with some precision: over-all noise level, composition of the noise, duration and distribution of exposure, and total time of exposure. These inquiries have led quite recently to a series of Federal and State laws that specify permissible noise conditions and prescribe compensation for workers suffering hearing loss due to occupational noise exposure.

In recognition of these possible effects of noise, more and more companies are establishing their own in-plant hearing test centers. When fully operational, these facilities will be used to screen individuals before they enter the working environment and at various time intervals during their occupational careers. In this manner, both the worker and the employer can be assured of mutual protection against the undesirable effects of noise pollution.

The G-S 1703 is a pure-tone audiometer with an intensity range of −10 dB to 90 dB HTL. It is extremely simple to operate, having only three operator pushbuttons – Start, Stop, and Hold. Included as an integral part of the unit is a recorder that makes a permanent record of subject responses, first for the left and then for the right ear. At each of the seven discrete frequencies presented, the subject's threshold is determined and recorded via a modified Békésy technique.

The G-S 1703 has at least two unique features not incorporated in other units currently available. First, the subject-controlled intensity changes at a variable rate, rapidly at the start of each test frequency, then more slowly as threshold is approached. This technique means that less time is spent getting to the threshold region at each frequency and more time is spent defining the threshold precisely. This, in turn, means greater retest reliability and a more meaningful audiogram.

Second, the G-S 1703 automatically initiates a check of threshold at 1 kHz at the end of each test. This value, when compared to the previous threshold value of 1 kHz, gives the operator an immediate indication of the validity of the test.

CONCLUSIONS

These distinctive features of the G-S 1703, added to its ease of operation and its reliability of design, should give it, like its more sophisticated antecedent, the G-S 1701, a long and healthy life in a world which increasingly requires precise information about the human hearing function. In conjunction with GR's growing line of instruments for sound measurement, these two units and their companions provide one of the most comprehensive single sources for audiometric and acoustic equipment.

–R. L. Grason

The author acknowledges, with gratitude, the work performed by Carol W. Hetzel in assembling and coordinating much of the material in this article.

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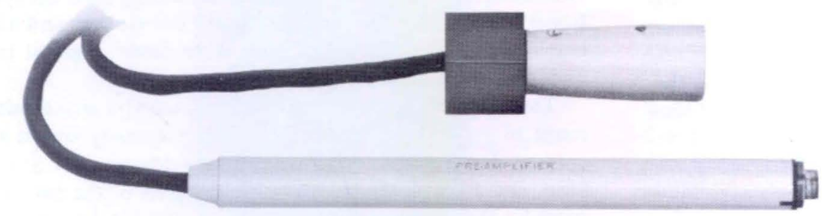
Condensed specifications for the G-S 1701 and 1703 Audiometers appear elsewhere in this issue.

Catalog Number	Description
	1701 Manual Diagnostic Audiometer,
1701-9700	117/234 V, 50 to 60 Hz
	1701 Automatic Diagnostic Audiometer,
1701-9710	117 V, 50 to 60 Hz
1701-9720	234 V, 50 to 60 Hz
1703-9700	**1703 Recording Audiometer**

A BIG LITTLE-BROTHER PREAMPLIFIER



GR 1560-P42 Preamplifier

1-in. microphone

1/2-in. microphone

1/4-in. microphone

1/8-in. microphone

The GR 1560-P42 Preamplifier is a bridge between most test microphones (ceramic or condenser) or ceramic transducers and the GR analyzers and sound-level meters. It is similar to the GR 1560-P40 unit but incorporates several improvements in its design.

Some Details

The -P42 unit is of smaller physical size (1/2-inch diameter by 6-inch length); incorporates switch-selected polarizing voltages, derived from an internal 65-kHz (approximate) oscillator, for condenser microphones; and has larger output current. Its three-wire output transmission system has a separate signal ground and a shield that does not carry signal current, thereby reducing hum pickup.

The standard front-end connection is readily adaptable to most test-measurement condenser and ceramic microphones. The input connection is guarded by a signal-driven shield, which reduces capacitive loading for low-capacitance microphones. Provision has been made, as an integral part of the preamplifier output jack (Figure 1), for insert-voltage calibrations, typically required for laboratory standard microphones such as the WE 640AA, and for remote checks of systems.

The preamplifier class AB output stage can provide up to 10 mA peak and > 1 V rms to feed full audio-range signals through cables as long as one mile; with no signal it draws less than 1 mA at +15 V, when used with ceramic microphones, thereby promoting longer supply battery life. Gain of the -P42 is switch-selected as unity (0 dB) or x 10 (20 dB). The FET input-stage design provides diode protection against input surges.

Connection between the preamplifier and transducers is by means of the accepted 0.460-60 thread, to fit present condenser microphones and their adaptors. Most other microphones and accelerometers are connected by use of simple GR adaptors.

Power for the -P42 unit is available from most of GR's sound analyzers and sound-level meters. For use with other instruments or for long cable runs, the GR 1560-P62 Power Supply will provide the required power. The power supply includes NiCad batteries, charging circuitry, an automatic low battery-voltage sensor to prevent excessive discharge, load-current limiting, and remote-switch control capability to turn off the power.

Other Necessities

Since no single type of microphone satisfies all test requirements, GR has made available as sets a group of microphones to supplement the preamplifier unit. The 1-inch ceramic and 1/2-inch condenser microphones are useful for measurements of low or moderate sound-pressure levels at normal audio frequencies. The 1/2-, 1/4-, and 1/8-inch condenser microphones are required when measurements are made at high sound-pressure levels or high frequencies. Each set includes all necessary adaptors to mate microphone, preamplifier, and GR 1562 Sound-Level Calibrator. In addition, an adaptor is available to mate the -P42 unit and standard 1-inch condenser microphones such as the Western Electric 640AA; another adaptor is supplied to mate to Switchcraft-type A3 audio-type connectors.

The GR 1560-9580 Tripod accommodates both the -P42 and -P40 preamplifiers, the GR 1560-P5 microphone, and all sound and vibration instruments having a 1/4-20 threaded tripod mount.

Development of the GR 1560-P42 was by E. R. Marteney.

Complete details of the GR 1560-P42 Preamplifier, microphones, and adaptors are available in GR Catalog U.

Catalog Number	Description
1560-9642	**1560-P42 Preamplifier**
	Ceramic Microphone Set
1560-9531	1-inch
	Condenser Microphone Sets
1560-9532	1/2 inch
1560-9533	1/2 inch
1560-9534	1/4 inch
1560-9535	1/4 inch
1560-9536	1/8 inch

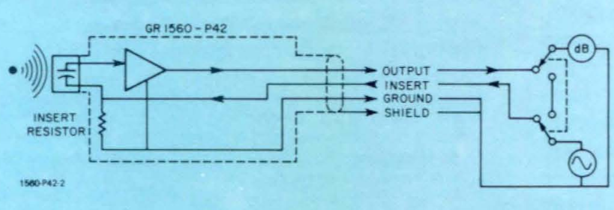


Figure 1. Schematic of 3-wire and insert-voltage connections in -P42 preamplifier.

Turn To Page 19 To Learn About Reduction Of Wind Noises

The Greeks Had A "Word" For It - Stroboscope

A coined word of deep impact in the field of illumination, derived from strobos (whirling) and skopeo (I look at)

GR 1542

Years ago man discovered that the eye could perceive rapidly moving objects by observing them only intermittently, as through a slit in a whirling opaque disc. A similar "miraculous" phenomenon familiar to early western-movie fans was commonly termed "wagon-wheel effect" and was due to the harmonic relationship between the information-sampling (camera-framing) rate and the rotational rates of the thousands of wagon wheels that thundered across the silver screen. But it took the introduction of Dr. Harold Edgerton's electronic stroboscope by General Radio Company some 38 years ago to bridge the gap between a novelty principle and a widely-useful tool. Intense microsecond flashes of light produced by the many improved general-purpose instruments introduced over the years provide the ultimate in motion-stopping capability to observe visually, to measure the speed of, or to photograph events and objects that would otherwise be but a blur.

Customer requests indicated a need for several additions to the GR strobe line such as the GR 1540 Strobolume® electronic stroboscope,[1] which produces extremely high light output, and the GR 1541 Multiflash Generator,[2] an accessory for photographic applications. New, less-expensive light generators also were requested for "single-use" applications that demand less versatile light sources, such as a simple basic stroboscope designed only to "freeze" motion for visual study. One answer to these needs is the GR 1542 Electronic Stroboscope.

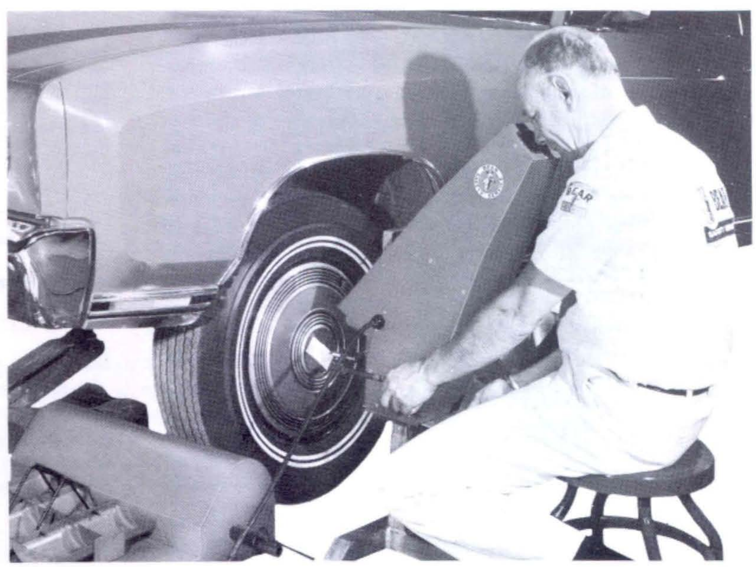
Small But Mighty

Simplicity in itself, the GR 1542 is a small, rugged, inexpensive and simple-to-operate stroboscope that puts the magic of frozen motion at your fingertips. Its stable wide-range oscillator produces steady stopped- or slow-motion images.

Most stroboscopes produce an image that appears to the operator to decrease in brightness as the flash rate is decreased. Usually this undesirable situation is overcome by switching additional capacitors into the discharge circuit at reduced oscillator frequencies. The GR 1542, however, uses a novel electronic circuit to provide an essentially constant sub-

[1] Miller, C. E., "Detailed Viewing in Ambient Brightness," *GR Experimenter*, September/October 1969.

[2] Miller, C. E., and Rogers, W. F., "New Shoes for an Old Workhorse," *GR Experimenter*, July/September 1970.



The Bear Corporation's portable automobile wheel-balancing machine incorporates GR 1542 stroboscope.

Photograph courtesy of Bear Corporation, Rock Island, Illinois.

C. E. Miller was graduated from Yale University in 1960 with a B. Eng. degree and received his MS degree from Massachusetts Institute of Technology in 1966. He joined General Radio in 1960 and is an engineer in the Component and Network Testing Group. He is a member of IEEE, AOA and ATI and holds a patent for a constant offset frequency-generating device to produce slow-motion images.

jective image brightness over a wide speed range **without switching**, resulting in greater operational simplicity and lower over-all cost!

In education, the GR 1542 is well suited to student use in experiments that demonstrate the principles of stroboscopy and harmonic motions. Industrial uses for the GR 1542 abound in development, test, and maintenance areas. Careful electrical and mechanical design assure that this hand-held instrument will perform faithfully under severe industrial environmental conditions.

A particularly interesting application involves the use of a slightly modified GR 1542 in a new-type automobile wheel-balancing machine (Figure 1) invented by Bear Manufacturing Corporation of Rock Island, Illinois – long the foremost manufacturer of automotive brake and wheel alignment equipment in the USA. The stroboscope is an essential component in the "Telabalancer." With this machine the operator quickly makes extremely accurate balances of wheels with all parts, including the hubcaps, in place and at road speeds of **120 mph!** Bear chose to use the GR 1542 for the critical illuminator in the Telabalancer because of its high, relatively-constant light output, oscillator stability, and ease of adjustment – important factors to the operator often working rapidly in areas of high ambient illumination. High reliability is also a prime requirement as the Telabalancer must function day after day. The stroboscope package provides the rather unique advantage that the entire strobe is bolted into the Telabalancer as a completely enclosed component. In the event service is required, the user simply unbolts and easily removes the stroboscope and returns it to his distributor for an exchange unit.

Some Background

Realization of the GR 1542 involved an unusual degree of cooperation among the industrial design, mechanical, and electrical engineers. The package was to be compact, neat in appearance, and the controls convenient to operate. It is physically rugged and thermally and electrically compatible with the high-performance strobe "innards," assuring operator safety and low cost. These criteria were met by use of a variety of tough plastics and a novel, symmetrical "clam-shell" injection-molded case, amazingly resistant to physical abuse.

A single-range uncalibrated oscillator provides flashing rates from approximately 180 to 3800 flashes-per-minute, or a speed ratio of approximately 20 to 1. Stability of the oscillator is of prime importance to produce steady, stopped, or slow-motion images. Accuracy of calibration, of course, is not a factor since the instrument is not intended for speed measurements. A five-turn continuous control provides vernier action for smooth speed adjustment throughout the oscillator range, and motions to above 50,000 rpm may be viewed by use of harmonics.

User response indicates the GR 1542 provides good performance and a truly economical capability to "stop" high-speed motions for visual study or analysis.

–C. E. Miller

The GR 1542 was designed by the author; W. A. Montague and P. A. d'Entremont contributed to the mechanical and industrial designs respectively.

Flash Rate: ≈ 180 to 3800 flashes per minute (3 to 63 flashes per second), continuously adjustable by uncalibrated 5-turn control.
Flash Duration: ≈ 5 μs at 63 fps, ≈ 25 μs at 3 fps.
Beam Angle: ≈ 40° at half-intensity points.
Power: 105 to 125 V, 50 to 60 Hz, 9 W.
Mechanical: Molded plastic case with face plate to protect lamp, high brightness reflector, standard 0.25-20 threaded hole for tripod mounting. DIMENSIONS (w x h x d): 4.2 x 2.16 x 7.52 in. (107 x 55 x 191 mm). WEIGHT: 1.8 lb (0.8 kg) net, 2 lb (0.9 kg) shipping.

Catalog Number	Description
1542-9700	**1542 Electronic Stroboscope**
1530-9400	**Replacement Strobotron Flash Lamp**

Information Retrieval

An article edited by L. J. Chamberlain, Executive Vice-President of Time/Data (a GR company), presents the subject of time-series analysis in comparatively simple terms and illustrations. It is entitled "A Simple Discussion of Time-Series Analysis" and is available to any reader who would like a single copy. Address your request to the Editor – *GR Experimenter,* General Radio Co., Concord, Mass. 01742.

EXTRA!

Stability in Standard Capacitors, Precision in Capacitance Measurements

Our standard capacitors **do** change with time.

Even though these capacitors have remained satisfactorily stable over the years, the accuracy of the calibrations, Figure 1, shows an inclination to plunge into the region where the uncertainties are less than a part per million. The National Bureau of Standards has led the way with improved capacitors and measurements. We have followed them, not only with interest but with some new instruments[1] designed to bring to other laboratories the increasing accuracy of NBS calibrations.

The new GR 1408 10- and 100-pF reference-standard capacitors with the proven mechanical and electrical stability of a fused-silica dielectric have been constructed to provide higher accuracy in the transfer and storage of the unit of capacitance. The new GR 1616 transformer-ratio-arm bridge has also been built to meet the need for improved precision in the intercomparison of these standards and for high accuracy in the calibration and measurement of a wide range of other capacitors. This bridge provides extended ranges of capacitance and conductance and extended sensitivity, particularly when used with the complementary new GR 1238 phase-sensitive detector and GR 1316 power oscillator. The bridge, detector, and oscillator assembly (GR 1621) is of value not only in the calibration of standards but also in the investigation of the dielectric properties of materials, through measurements of small capacitances and conductances and of very small changes in these quantities.

[1] Abenaim, D. and Hersh, J. F., "New Fused-Silica-Dielectric 10- and 100-pF Capacitors and a System for Their Measurement," *IEEE Transactions on Instrumentation and Measurement*, November, 1970.

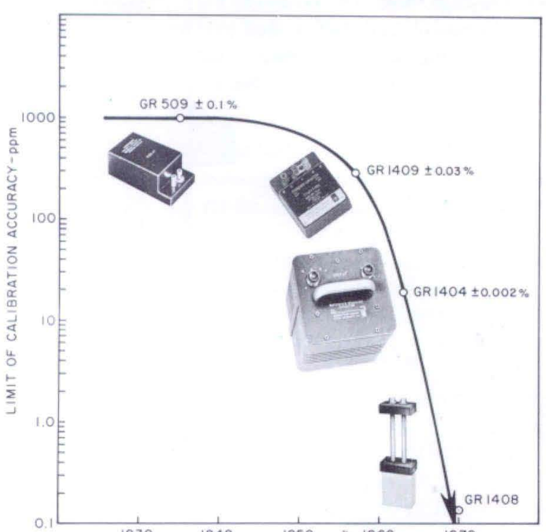
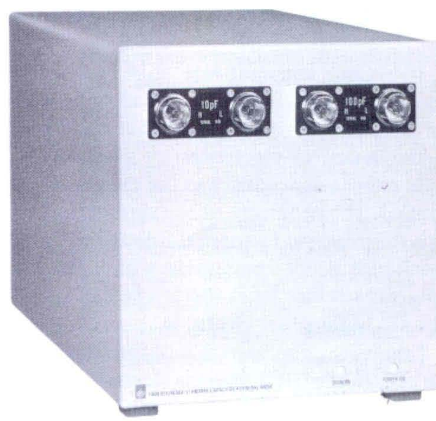


Figure 1. GR standard-capacitor calibration-accuracy improvement with time.



GR 1408 standard, with temperature-controlled air bath.

FUSED-SILICA CAPACITORS GR 1408 REFERENCE STANDARD CAPACITORS

The fundamental design of the new 10- and 100-pF standards is based upon the development at NBS by Cutkosky and Lee[2] of a 10-pF capacitor with time stability and small variations due to voltage change or shock, which permit calibration to parts in 10^7. The dielectric material used for such stability is a special grade of fused silica. It has the further advantages of low losses and low frequency dependence of its dielectric constant in the audio frequency range.

GR Design

The two main considerations in the design of the fused-silica capacitors were the manner of applying the electrodes to the substrate and the manner of supporting the capacitor

[2] Cutkosky, R. D. and Lee, L.H., "Improved Ten-Picofarad Fused-Silica Dielectric Capacitor," *NBS Journal of Research*, Vol. 69C, July-September, 1965.

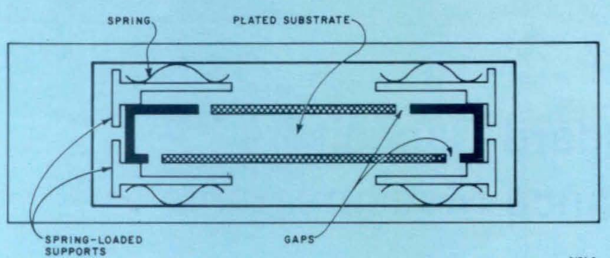


Figure 2. Substrate thickness
is the only difference between 10- and 100-pF units.

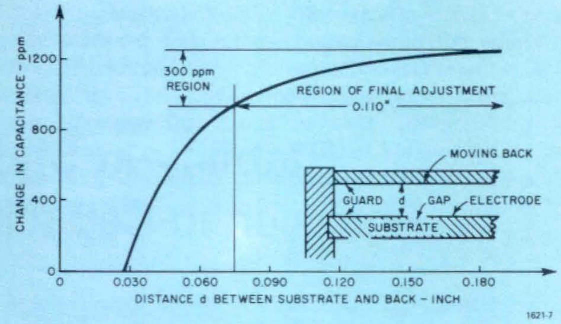


Figure 3. Effect upon capacitance
of changes in back-substrate separation.

in its cell. It became apparent very quickly that the gap between these electrodes was critical. It had to be well defined and free of isolated particles of metal that could be attracted to the plated guard or electrodes by electrostatic forces, which would cause a dependence of the capacitance upon the voltage applied. The geometry of the support in the vicinity of the gap is the principal factor in the design of the cell as the direct capacitance is not completely within the fused silica but includes capacitance from the top face of one electrode through the gap to the other electrode.

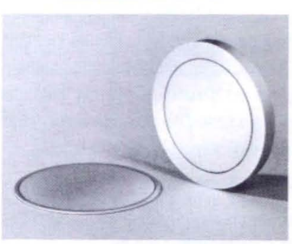
Figure 2 shows the configuration of the electrodes and the supporting cell. In the GR design the electrodes and guard are in the same plane on each face, and photo-etching techniques can be used both to generate the gaps and to adjust the capacitance by changing the area of an electrode. Fortunately, microelectronic techniques are available at GR for the deposition of electrodes and the generation of gaps. These techniques have proven more predictable and reliable than any grinding or masking technique investigated. Since the electrode areas are not equal and the capacitance is defined mostly by the area of the smaller electrode, only one gap is now crucial.

The substrate is held between spring-loaded supports which are shaped so that, even if the substrate moves, the guard in the vicinity of the gap stays the same. As the distance between the plane of the gap and the holder above it changes so does the direct capacitance. Figure 3 shows the magnitude of this effect. It appears from this graph that it is "unhealthy" to have the guard too close to the gap but, as the distance becomes greater than about 70 mils, the position of the guard is less critical. We take advantage of the last 300 ppm of change to provide for motion of the guard in the cell as a final capacitance adjustment.

Construction

Figure 4 shows two coated and etched capacitor substrates. The coating consists of 0.0005 inch of pure gold. Both the thin substrate (0.030 inch thick) for 100 pF and the thick one (0.300 inch thick) for 10 pF have a diameter of 2.727 inches. The element is placed in a brass holder, and the capacitance is adjusted to ±100 ppm of nominal values. Contact to the electrodes is made through gold-coated phosphor-bronze springs. Figure 5 shows the holder ready to be placed in a stainless-steel container, and also shows the assembled and sealed cell. This container is welded shut, evacuated,

Figure 4. 100-pF and 10-pF
substrate elements.



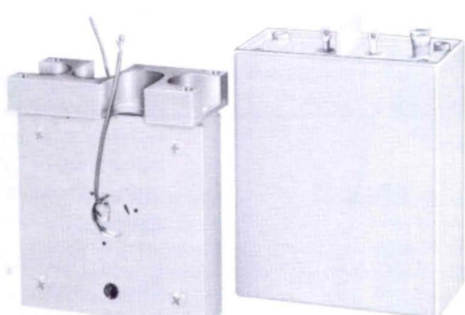


Figure 5. Capacitor brass holder
and assembled and sealed cell.

Figure 6. Oil-bath
version of GR 1408 capacitor.

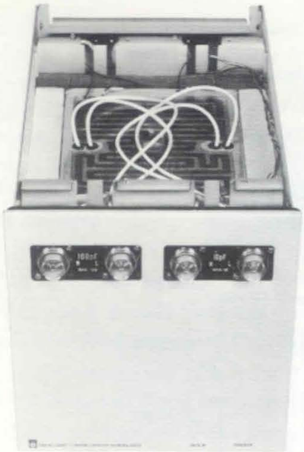


Figure 7. Air-bath version of GR 1408 capacitor.

baked, back-filled with dry nitrogen, and sealed; connections to the capacitor are made via glass-to-metal feedthroughs.

The dielectric constant of fused silica has a temperature coefficient of approximately 10 ppm/°C. To make meaningful measurements at a level of a part in 10^7, one has to know the ambient temperature to within 0.01°C. This can be accomplished in an oil bath; Figure 6 shows the oil-bath version of the capacitor. The GR 874® connectors, gold plated for lower contact resistance, are installed six inches above the capacitor to allow connection above the oil level. The normally simple measurement of this capacitor in an adequate oil bath is, however, complicated by the additional precision apparatus required to make the accurate temperature measurements needed to define the capacitance value. For that reason, an air bath was developed which can provide one or two capacitors with their own environment and, therefore, eliminate temperature measurements except in the case of highest accuracy. The air bath, Figure 7, is thermostatically controlled at a nominal 30°C. The bath has a long-term stability of 0.01°C; it changes by less than 0.01°C for a 6°C change in ambient temperature. Temperature control is by a 12-volt system; batteries can be used during transportation.

Performance

Evaluation of the fused-silica capacitors was difficult because standards and measuring equipment capable of the required accuracy and resolution were not available. We developed this equipment concurrently with the capacitors.

We found the voltage dependence of a fused-silica capacitor to be a good indicator of its quality, and it is the first test made on all new units. The capacitance change has to be less than a part in 10^8 when the voltage applied is changed from 50 to 150 V.

Frequency dependence and dissipation factor are directly related to the dielectric and were evaluated by comparison with two types of air capacitors. Our tests showed the frequency dependence to be a few ppm between 1 and 10 kHz; the dissipation factor was also a few ppm at 1 kHz. Measurement accuracy was 3 to 4 ppm.

The effects of mechanical and thermal shocks were investigated. Oil-bath versions were dropped at different angles: the capacitance did not change by more than a part in 10^7. Some 100-pF assemblies showed more shock sensitivity (1 ppm), attributed to bowing of the thin substrate. The capacitors were also cycled between 0 and 50°C and the hysterisis effect was less than 4 ppm; the cause of this change is questionable but could be due to temperature-measurement uncertainties of our oil bath.

As for long-term stability, it could only be checked indirectly. The difference between two capacitors in an air bath did not change by more than one part in 10^7 during a one-year observation period.

GR 1621
PRECISION CAPACITANCE-MEASUREMENT SYSTEM

One consequence of the improved quality of the new capacitors is that both the tests required to demonstrate their stability and the calibrations to be made in their ultimate use as standards require more precision than that to be found in the measurement systems of our laboratory and, indeed, of most laboratories. We met this need by developing the GR 1621 Precision Capacitance-Measurement System (Figure 8). The system is comprised of the GR 1616 Precision Capacitance Bridge, 1316 Oscillator, and 1238 Detector. A specific design objective of this system was to provide precision in intercomparison measurements of our new capacitors to parts in 10^8 near 1 kHz. An equally important design objective was to provide direct readings with high accuracy in measurements of a wide range of capacitance and conductance at audio frequencies.

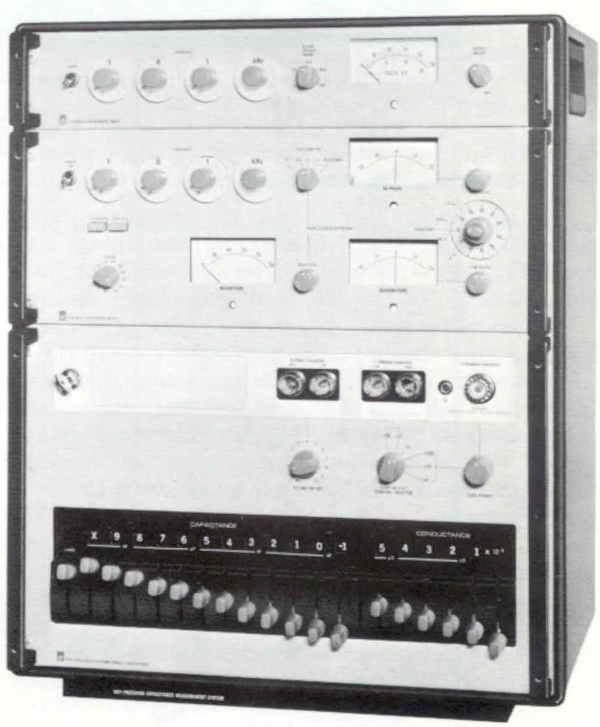


Figure 8. GR Precision Capacitance-Measurement System. Top to bottom: 1316 Oscillator, 1238 Detector, 1616 Precision Capacitance Bridge

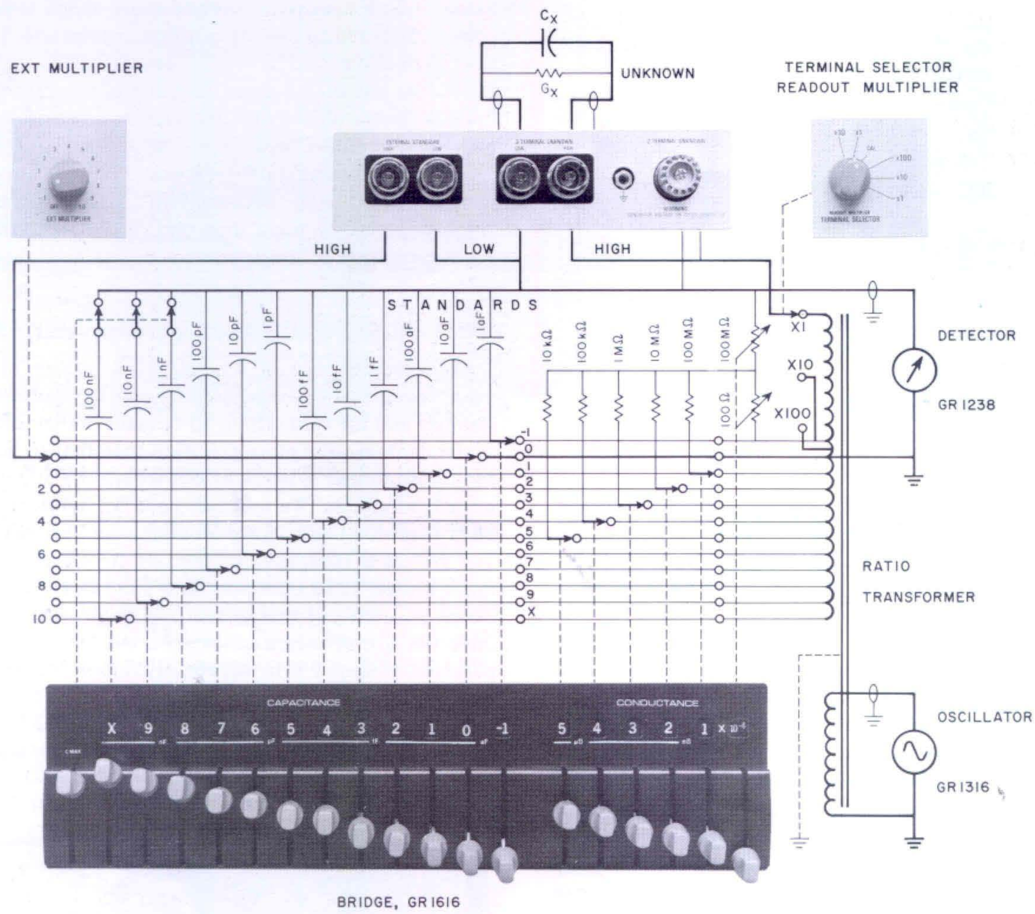


Figure 9. GR 1616 bridge: Basic circuits.

GR 1616 Precision Capacitance Bridge

The new bridge (Figure 9) uses the familiar transformer-ratio-arm bridge circuit.[3] Greater precision through higher applied voltages – 150 volts at 1 kHz – is obtained by use of a three-winding, 200-turns-per-winding toroidal transformer.

The bridge has twelve decades of capacitance provided by twelve internal standard capacitors ranging from 100 nF to 1 aF, and by the eleven taps on the transformer winding that give decade steps from 10 through 0 to –1. The three highest value capacitors can be disconnected in sequence when not needed, with a consequent reduction in detector shunt-capacitance loading and increase in bridge sensitivity. For the stability required in precision intercomparisons, the eight highest value capacitance standards are sufficiently insulated thermally to provide a time constant of at least six hours, for changes in the ambient temperature of the bridge.

Losses in the unknown capacitors are balanced by conductance decades, by use of five internal conductance standards – three metal-film precision resistors (10 kΩ, 100 kΩ, and 1 MΩ) and two carbon-film resistors (10 MΩ and 100 MΩ).

[3] Hersh, J. F., "Accuracy, Precision, and Convenience for Capacitance Measurements," *The General Radio Experimenter*, August-September, 1962.

Some operational features are:

- Capacitance measurement range – 10 μF to 0.1 aF (10^{-5} to 10^{-19} F).
- Limits of capacitance measurement errors range from 10 ppm (1 nF, 100 pF, 10 pF standards) to 50 ppm from 1 kHz to below 100 Hz, measured at 23° ±1°C.
- Conductance measurement range – 10^{3} to 10^{-10} μmhos.
- Limit of conductance measurement error is 0.1% of reading at 1 kHz, over most of the range.
- Gold-plated GR874 coaxial connectors for low and repeatable contact resistance; GR900® connector for coaxial capacitance measurements.
- Terminals available for external-standard use.

GR 1238 Detector and 1316 Oscillator

The output of the GR 1616 bridge is only a few hundredths of a microvolt when the input to the bridge is 100 volts and the unbalance is a part in 10^{8} of 10 pF. The new detector developed to extract this small signal from noise, at the high impedance level of the bridge output, is a combination of a high-impedance, low-noise preamplifier, a tuned amplifier with 130-dB gain, and two phase-sensitive detector circuits. The input impedance is that of 1 GΩ in parallel with 20 pF; the noise voltage at 1 kHz with a source impedance equivalent to the output impedance of the bridge in the measurement of 10 pF, i.e., 100 MΩ in parallel with 500 pF, is about 30 nV per root Hz.

A resumé of operating features of the oscillator and detector includes:

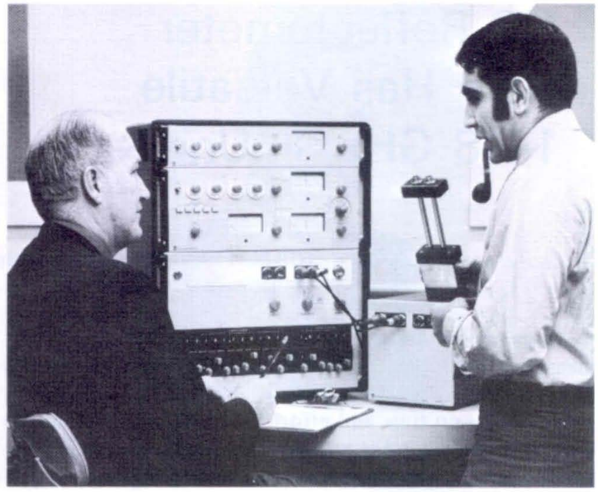
- Detector input is protected by diodes to the fullest extent of the oscillator's output of 150 volts.
- Matched decade switches tune the detector and set the oscillator frequency over the range 10 Hz to 100 kHz.
- A line-frequency notch filter is available in the detector, plus a choice of linear or compressed meter response.
- Detector bandwidth is narrowed by adjustment of the integration time constant from 0.1 to 10 seconds, to reduce noise.
- Bridge balance is speeded by the presence of detector panel meters that display magnitude of the unbalance signal plus the in-phase and quadrature components. The two phase meters can be made to respond to capacitance and conductance independently by adjustment of phase relationships. The phase reference voltages can be rotated from 0 to 360° to achieve any phase condition.
- The oscillator provides two 90°-displaced fixed-voltage reference signals to the phase sensitive detectors.
- Oscillator output is readily controlled and monitored by a 5-position range switch, vernier control, and panel meter.
- Oscillator signal distortion is typically less than 0.3% with loads ranging from short to open circuit.
- Amplifier output and meter outputs are available on the rear panel.

CONCLUSION

The new reference standard 10- and 100-pF capacitors, together with the new measurement system, make it easy for standards laboratories to improve their accuracy in the transfer and storage of the unit of capacitance. The new measurement system will also provide good direct-reading accuracy, to 10 ppm under limited conditions, in the calibration of a wide range of capacitors from 10 μF to much less than a picofarad, at audio frequencies. From our experience eight years ago with the introduction of new standards (GR 1404 capacitors) and a new measurement system (GR 1620), we can make two predictions: (1) Although this extended resolution will solve some measurement problems, it will also reveal some new ones when we try to make the sixth, seventh, or eighth figure significant. (2) Although we have provided more resolution in the system than most of us need today, we know you will be asking for more resolution tomorrow.

– D. Abenaim
J. F. Hersh

Condensed specifications for the GR 1408 Reference Standard Capacitor and the GR 1621 Precision Capacitance-Measurement System appear elsewhere in this issue.



D. Abenaim ***(right)*** graduated from George Washington University in 1965 (BSEE) and received the MSEE degree in 1967 from GWU. He joined GR in 1965 as a development engineer, principally concerned with standards and precision measurements. During a leave of absence in 1967, Dan was at the National Bureau of Standards, working with transportable voltage standards and fused-silica capacitors. He is a member of Tau Beta Pi and IEEE.

J. F. Hersh ***(left)*** graduated from Oberlin College with the AB degree (1941) and went on to Harvard University for his MA in Physics and PhD in Applied Physics (1942 and 1957). His doctorate work was in the field of electromechanical transducers. John's experience has involved work at Harvard's Underwater Sound Laboratory and teaching physics at Wellesley College. He joined GR in 1957 as a development engineer but shared time with the National Bureau of Standards that year, involved in bridge and capacitance research. His work since has been with inductance and capacitance bridges and the development of improved standards and techniques. He is a member of IEEE, Phi Beta Kappa, and Sigma Xi.

Catalog Number	Description
	1621 Precision Capacitance-Measurement System
1621-9701	Bench Model
1621-9702	Rack Model
	1616 Precision Capacitance Bridge
1616-9700	Bench Model
1616-9701	Rack Model
	1316 Oscillator
1316-9700	Bench Model
1316-9701	Rack Model
	1238 Detector
1238-9700	Bench Model
1238-9701	Rack Model
	Reference Standard Capacitor, air bath
1408-9700	**1408,** 10 pF
1408-9702	**1408,** 10/10 pF
1408-9703	**1408,** 100 pF
1408-9705	**1408,** 100/100 pF
1408-9706	**1408,** 10/100 pF
	Reference Standard Capacitor, oil bath
1408-9701	**1408-A,** 10 pF
1408-9704	**1408-B,** 100 pF

GR Reflectometer Now Has Versatile 1–18 GHz RF Unit

GR 1641 Sweep Frequency Reflectometer with the 18-GHz RF Unit.

There are definite advantages in providing the facility for measuring the reflection (SWR) and transmission properties of networks over the widest possible frequency bandwidth. First, the set-up time and effort for measuring components that have differing band-center frequencies are greatly reduced. Second, the equipment cost is lower when one high-directivity directional-coupler assembly can be employed in place of a number of octave-band couplers (in this case, about five).

It is, furthermore, advantageous to have the best possible directivity in a network-analyzer directional coupler because it means making a direct measurement without the need for computer correction. An accurate, continuous sweep-frequency measurement can be performed in contradistinction to the step-frequency measurement required for computer-correction.

The GR 1641-9603 RF Unit comes closer to meeting these and other requirements than any of its predecessors.

A Review of the System

The complete GR 1641 Sweep-Frequency Reflectometer,[1] a type of network analyzer, has distinct features that make it ideal for the measurement of microwave components. Its original concept was to provide the simplest, easiest-to-use instrument for measurement of the magnitude only of reflection coefficient (return loss or SWR) and transmission coefficient (insertion loss) of networks or microwave devices. If there is no need to measure the phase of these parameters, then considerable simplification of the set-up and operation of the measuring instrument results. The earlier GR 1641 offers this simplification with the result that, after a simple initial level-set adjustment, the instrument is completely calibrated and ready for use. The adjustment is stable; there is no perceptible drift. This approach necessitates the inclusion of all the directional-coupler "plumbing" within the package, and this alone offers an advantage to the user. He is not required to gather up a hodge-podge of couplers, detectors, and cables to make up a measurement system.

Some users, however, may prefer not to be bound by this concept. The fact is that the individual main-frame and plug-in units of the GR 1641 can be operated in other measurement systems. In particular the new GR 1641-9603 RF Unit, which covers the frequency range from 1 to 18 GHz, can be used in *any* network-analyzer to take advantage of the exceptional directivity, the wide bandwidth, and the 18-GHz operating frequency of this unique reflection-measuring device.

Alternately, the GR 1641 Main Frame and Indicator may be employed with whatever plumbing the user chooses. In fact, if transmission or insertion-loss measurements only are required, this plumbing is nothing more than attenuator pads and a detector. An example is the GR 1641-P3 Transfer Detector. With this unit, insertion loss as high as 60 dB can be measured.

The New 1-18 GHz RF Unit

The 1641-9603 RF Unit contains a newly-developed directional coupler that has unusually good directivity over a wide band. This coupler has directivity performance comparable to the best octave-band couplers. Also, the reflection coefficient or SWR looking back into the UNKNOWN terminals is quite low. Both these characteristics contribute to accuracy, a subject discussed below. The directivity is illustrated in Figure 1 in terms of both dB directivity and residual

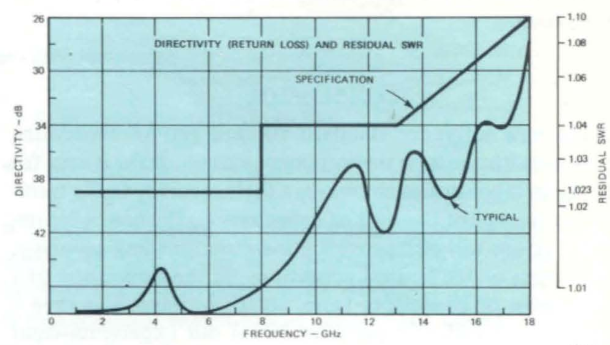


Figure 1. RF-unit directivity.

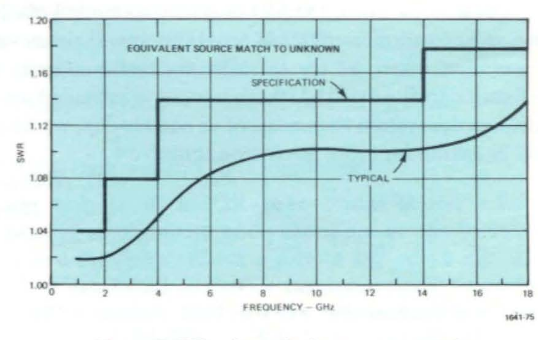


Figure 2. RF-unit equivalent source match.

[1] MacKenzie, T. E., et al, "The New Sweep-Frequency Reflectometer," *GR Experimenter*, March-April 1969.

SWR; the equivalent-source-match SWR is illustrated in Figure 2. The UNKNOWN connector is an IEEE Standard No. 287, GPC-7mm. The instrument may be converted easily to N, SMA, TNC and other popular connectors by means of precision adaptors.

The rf-unit block diagram is given in Figure 3. One of the two identical directional couplers is used to sample the source signal and normalize it by leveling, so that the incident wave to the UNKNOWN is maintained constant. The tracking of the two couplers is important for this. An envelope or "video" detector is employed to provide the leveling signal. The second coupler is used to measure the reflection from the device under test connected to the GPC-7mm connector. The reflected signal is detected in a second envelope detector in front of which is installed a low-SWR, constant-attenuation, 10-dB attenuator. This attenuator provides an improved match not achievable with broad-band diode detectors.

The detectors are affixed by means of GPC-7mm connectors and can be removed for other applications.

The couplers have a nominal coupling value of 19 dB, ±1 dB approximately, from 4 to 18 GHz. At 3 GHz the coupling is 21.5 dB, at 2 GHz it is 25.5 dB, and at 1 GHz it is 32 dB.

The 1-18 GHz Transfer Detector

The GR 1641-P3 Transfer Detector is a well-matched envelope-detector assembly comprising a 10-dB attenuator and a diode detector. The attenuator is employed to improve the match to the diode detector. With the GR 1641 indicator system the assembly has a sensitivity of −65 dBm. The SWR specification is $1.02 + 0.005 f_{GHz}$.

Accuracy

Although accuracy was described in detail on page 8 of the referenced *Experimenter*, a brief qualitative discussion is in order.

The significance of the directivity, Figure 1, and the source match, Figure 2, are illustrated in the following expression:

$$|\Gamma_i| = \Gamma_o + k\Gamma_x + k\Gamma_s\Gamma_x^2$$

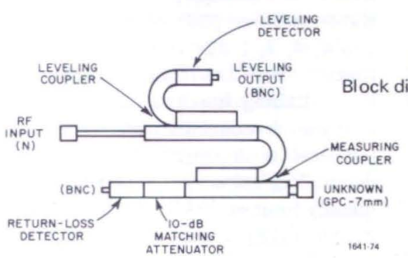


Figure 3. Block diagram of rf-unit.

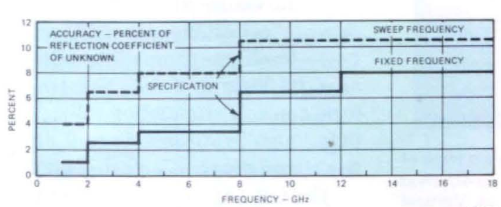
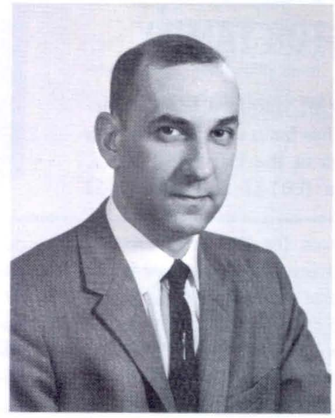


Figure 4. RF-unit accuracy.



John Zorzy received his BS-Physics degree from George Washington University in 1948 and his MS from Tufts College in 1950. His early experience was directly related to design and development of radar, antennas, and microwave devices. He joined GR in 1960 and presently is Group Leader of the Microwave Group. John is a member of IEEE, Sigma Xi, Sigma Pi Sigma, and serves as Chairman of the EIA/NCTA Task Group on 75Ω Precision Coaxial Connectors. He is a member also of the IEEE Subcommittee on Precision Connectors, of JEDEC Committee JS-9, and of the Department of Commerce Joint Industry Research Committee for Standardization of Miniature Precision Coaxial Connectors.

where:

Γ_i = Reflection coefficient indicated by the 1641 system.
k = Normalized frequency response of the 1641 system.
Γ_x = True, unknown reflection coefficient.
Γ_o = Residual "directivity" reflection coefficient (Figure 1).
Γ_s = Reflection coefficient looking into coupler (Figure 2).

When the UNKNOWN has low SWR ($\Gamma_x \simeq < 0.1$), the significant term is the directivity, Γ_o. When the UNKNOWN has moderate or high SWR, the significant terms are $k\Gamma_x$ and $k\Gamma_s\Gamma_x^2$. In this latter case, the specifications shown in Figure 4 are expressed as a percent of Γ_x for simplicity.

—J. Zorzy

Complete specifications for the GR 1641 are in the supplement to GR Catalog U, to be distributed shortly.

	Bench Models	Rack Models
1641 Sweep-Frequency Reflectometer		
20 MHz to 1.5 GHz	1641-9702	1641-9712
20 MHz to 7 GHz	1641-9701	1641-9711
20 MHz to 18 GHz*	1641-9704	1641-9714
20 MHz to 18 GHz	1641-9705	1641-9715
500 MHz to 7 GHz	1641-9703	1641-9713
500 MHz to 18 GHz	1641-9706	1641-9716
1 GHz to 18 GHz	1641-9707	1641-9717
1641-Z Sweep-Frequency Reflectometer, with display oscilloscope		
20 MHz to 1.5 GHz	1641-9902	1641-9912
20 MHz to 7 GHz	1641-9901	1641-9911
20 MHz to 18 GHz*	—	—
20 MHz to 18 GHz	—	—
500 MHz to 7 GHz	1641-9903	1641-9913
500 MHz to 18 GHz	—	—
1 GHz to 18 GHz	—	—
Transfer Detectors, included (where appropriate) with 1641 and 1641-Z; not included with RF Units purchased separately		
20 MHz to 7 GHz	1641-9606	
1 GHz to 18 GHz	1641-9604	
RF Units, to fill partially equipped models		
20 MHz to 1.5 GHz	1641-9601	
500 MHz to 7 GHz	1641-9602	
1 GHz to 18 GHz	1641-9603	
1641-9605 Accessory Kit	1641-9605	

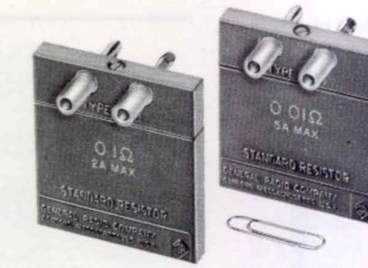
* Includes all three RF Units

other new products

EXPANSION IN THE RESISTOR FAMILY

The family of standard resistors designed by GR has been increased by two more members in the GR 1440 series. These are the 0.01-Ω and the 0.1-Ω resistors.

Principal uses for the new resistors are in calibrations of low-impedance systems, for use in substitution measurements, and as laboratory or production standards. Construction of the new resistors is somewhat different from resistors of higher values in the 1440



series. Previously, use was made of the card-type wire-wound technique. The new resistors are made up of a low-inductance meander-cut sheet element of well-aged Manganin,* connected to gold-plated copper terminals.

*Registered trademark of Driver-Harris Co.

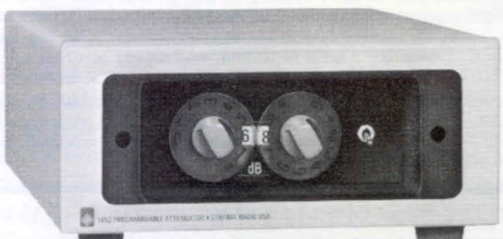
When completed, the resistors are adjusted, with relation to the nominal value, to 0.1% (0.01 Ω) and 0.05% (0.1 Ω) respectively. Both units are oil filled and sealed into oil-filled, diallylphthalate boxes for long-term stability and mechanical protection.

Development of these resistors was by W. J. Bastanier.

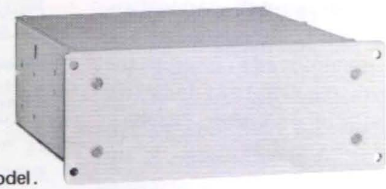
Complete specifications for the GR 1440 0.1-Ω and 0.01-Ω resistors are in the supplement to GR Catalog U, to be distributed shortly.

Catalog Number	Description
1440-9671	**1440 Standard Resistor,** 0.01 ohm
1440-9681	**1440 Standard Resistor,** 0.1 ohm

SOLID-STATE, PROGRAMMABLE ATTENUATORS



Manual-remote model.



Remote only model.

Automated testing is a must for efficient high-volume production. If manually operated electronic instruments are involved, they have to be designed so that controls can be set and changed remotely by suitable electrical signals, often under computer control. Calibrated programmable attenuators are useful for a variety of tasks

- to extend and/or program the dynamic range of other test equipment such as analog or digital meters, level sensors, oscilloscopes, wave or spectrum analyzers, and counters
- as gain or loss standards for measurements using insertion techniques
- for level setting of signal sources for receiver testing of sensitivity, over-load characteristics, and selectivity
- for open- or closed-loop leveling of sources responding to a preset program or a detector with suitable analog/digital conversion.

Note that signal sources often have provision for frequency programming but remote level setting is limited or unavailable.

Calibrated attenuators now available are almost all electromechanically programmed. Relays, reed switches, and turrets of coaxial pads operated by motors or solenoids are in use. With this approach, excellent results in terms of accuracy, SWR, and broad coverage can be obtained, but switching time suffers and there are definite limits to operating life.

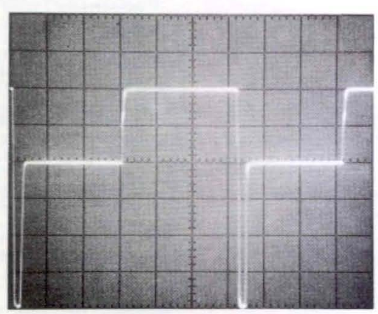


Figure 1. Programming pulse triggers trace of attenuation transition from 0 dB to 40 dB at 30 MHz. Horizontal scale: 1 ms/cm. Vertical scale: 10 dB/cm.

These are severe restrictions which would rule out applications in highly repetitive production testing or high-speed systems. The GR 1452 Attenuator was designed specifically for these requirements. Since it is all solid state, it is fast and not subject to the usual performance-life limitations. One-dB steps from 1 to 80 dB are produced by highly stable resistive pads of 40, 20, 10 dB and 8, 4, 2, 1 dB. These T and π-pads, together with the rf diodes performing the switching function, are assembled into pseudo-coaxial structures. Switching time is dependent largely on the decoupling networks; with a lower frequency limit of 10 kHz it is less than 0.5 ms. Operating frequencies are 10 kHz to 500 MHz.

As the choice of pad values suggests, remote selection uses BCD coding. Logic levels are compatible with common IC logic systems. Negative true logic controls attenuation values (all inputs "high" corresponds to 0 dB relative attenuation).

The GR 1452 is available in two versions. One model, for remote operation, is not packaged in an instrument case to save money and space. The second model offers both manual and remote operation (when the dials are set to the "R" position).

Figure 1 shows a transition in attenuation from 0 dB to 40 dB, at 30 MHz. A spectrum analyzer was used as the detector (with suitably wide bandwidth and not scanning frequency); the horizontal sweep is triggered by the switching signal applied to the attenuator.

Development of the GR 1452 was by G. H. Lohrer.

Complete specifications for the GR 1452 are available in the supplement to Catalog U, to be distributed shortly.

Catalog Number	Description
	1452 Programmable Attenuator
1452-9700	Manual-Remote Bench Model
1452-9701	Manual-Remote Rack Model
1452-9702	Remote-Only Model
	0480-9722 Adaptor Set, to rack-mount manual-remote model

WINDSCREENS FOR MICROPHONES

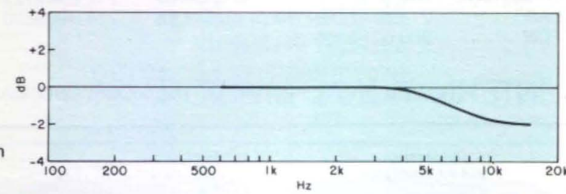


Figure 1. Effect of 1560-9521 Windscreen on response of 1560-P5 Ceramic Microphone.

The GR windscreens for 1-inch diameter microphones (1560-9521) and 1/2-inch diameter microphones (1560-9522) have been added to the ever-growing list of small accessories designed to make the lot of the sound-measurement engineer a little easier. The two new windscreens are fabricated of reticulated polyurethane foam. They are especially helpful if one must make noise measurements outdoors when winds are of low velocity (30 mph or lower). Wind-generated noise is attenuated by a factor of 20 dB or more.

These windscreens should prove to be useful also in industrial areas that may be oily or dusty. The use of a polyurethane windscreen will protect the microphone diaphragm and has only a very small effect on the microphone frequency response and sensitivity. As the windscreen becomes soiled it can easily be removed, washed, and reused.

The effect on the microphone sensitivity and frequency response is shown in Figure 1 for the GR 1560-9521 windscreen. Loss is essentially 0 dB below 3 kHz, then rises to about 2 dB at 12 kHz.

Development of the windscreens was by E. E. Gross, Jr.

Catalog No.	Description
	Windscreen, 4 per pack
1560-9521	1-inch diameter
1560-9522	½-inch diameter

A COUNTER IMPROVES

The GR 1192-B Counter is another answer from GR to consumer requests for economical instrumentation. It incorporates higher frequency response (50 MHz) than its predecessor, the 1192-A, without relaxing sensitivity specifications. With type-acceptance by the FCC as an a-m frequency monitor and as a monitor for fm and vhf-television broadcasts, the 1192-B (plus the 1157-B Scaler for approved operation to 216 MHz and to 500 MHz for normal tests) is available to broadcasters for use as a frequency monitor.

Obviously, we have no wish to restrict the use of our counter to the broadcast industry. Any reader interested in reading frequency, frequency ratios, time intervals, single and multiple periods, and many other electrical phenomena for which the electronic counter is suited, should refer to the article[1] that introduced the 1192-A to *Experimenter* readers. Many features of the 1192 counters are explained there in detail.

As in the previous model, data displays can be varied to suit test requirements; options are 5-, 6-, or 7-digit readout. Buffered data-output versions are available for each of the digit options.

Development of the GR 1192-B was by S. Bentzen.

Complete specifications for the GR 1192-B Counter are in the supplement to GR Catalog U, to be distributed shortly.

1192-B Counter (50 MHz) Bench Models
5-digit readout
6-digit readout
7-digit readout
1192-Z Counter (500 MHz with scaler) Bench Models
5-digit readout
6-digit readout
7-digit readout
Relay-rack mounting for 1192-B or 1192-Z
Option 2 BCD Data Output for 1192-B or 1192-Z
1158-9600 Probe, Tektronix P6006 (010-0127-000) not sold separately

[1] Bentzen, S., "The Counter Punch," *GR Experimenter*, July/August 1969.

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