THE
GENERAL RADIO

Experimenter

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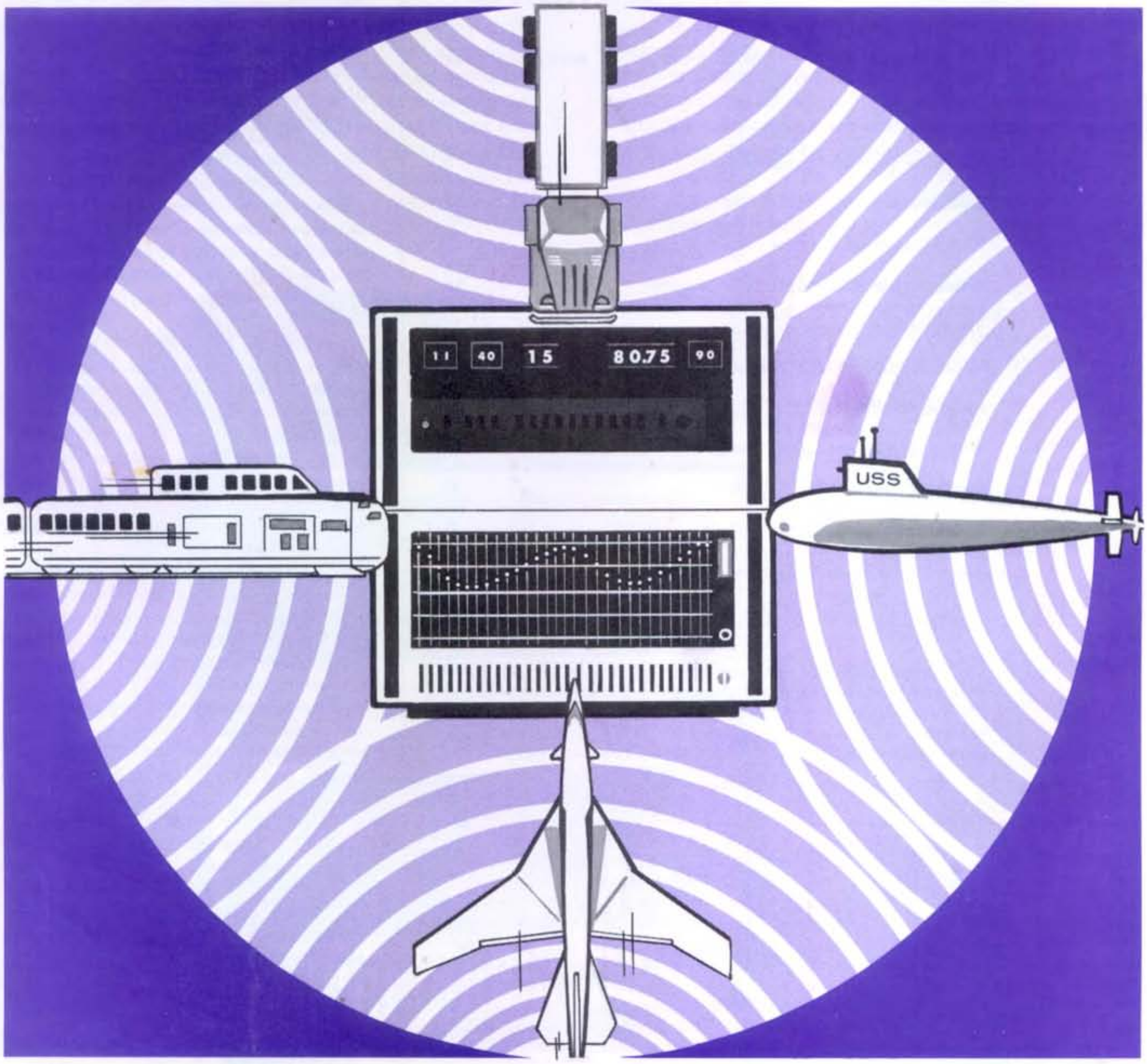

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The *General Radio Experimenter* is mailed without charge to engineers, scientists, technicians, educators, and others interested in the instruments and techniques of electrical and electronics measurements. Address all correspondence to Editor, *General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

THE COVER Experts in the field of sound and vibration are face to face with a problem peculiar to twentieth-century civilization – NOISE. We hear, from all directions, of the effects of noise pollution upon the well-being of the general public. We recognize also that noise can be an indicator of trouble, an element of danger, or a mask to cover desired information. In order to mitigate noise, we must evaluate its effects by performing measurements and studying the efficiency of noise-reduction efforts. Most desirable in noise studies is an ability to make instant evaluations through real-time measurements. Our cover depicts some areas of concern to the public, to engineers, and to the government. GR hopes that its latest contribution, the GR 1921 Real-Time Analyzer, will help establish new standards for allowable noise levels, improve mechanical designs, and contribute to the health and safety of all people.

A recent article in the National Conference of Standards Laboratories ***NCSL Newsletter*** asked the question "Do specifications really specify?" Editors of several trade journals also have been beating the drums to start a parade for clear and undeceptive specifications. A most important point raised was reflected in the comment ". . . bad specs drive out good ones."

The subject of specifications is one that is very sensitive to the economic feelings of instrument manufacturers. Specifications are the links that form the bond between producer and customer. That bond is only as strong as the weakest link, which could easily be a specification that is not attainable or can be met only under controlled (but not specified) circumstances.

A link can be found weak for other reasons, one of which is use of unclear or strange language. Lack of definitions for special terminology employed in some fields of instrumentation technology is, too often, the root of the evil of ambiguous wording. Another weakness may arise in the choice of tolerances; these ***can*** be based upon actual production test results with a clear understanding, or expression, of the standard deviations of the basic test data. Unfortunately, no standardized approach exists with which manufacturers may comply and the customer is left, too often, to determine how accurately a tolerance has been derived.

More thought is required, both of customer and producer, which will lead to a recognition of what is good, what is necessary, and what is sufficient in presenting specifications. General Radio has pondered the problem of specifications for many years and, in its own way, has tried to present to the customer what it believes is a true specification for instrument performance. We wonder sometimes whether we might say less, or more, in presenting a parameter for instrument performance. Perhaps our readers have some thoughts on the subject?

C E White

C. E. White
Editor

New-Generation Acoustical Analyzer

A sound and vibration analyzer, conceived as a third-generation answer to analysis and interpretation of an increasingly complex environment on land, under sea, and in the air. This prodigy gives 1/3-octave spectrum analysis from 3.15 Hz to 80 kHz and employs a unique digital-detection technique to achieve performance unattainable with analog instruments.

by W. R. Kundert, J. A. Lapointe, and G. R. Partridge

THE BROAD VIEW

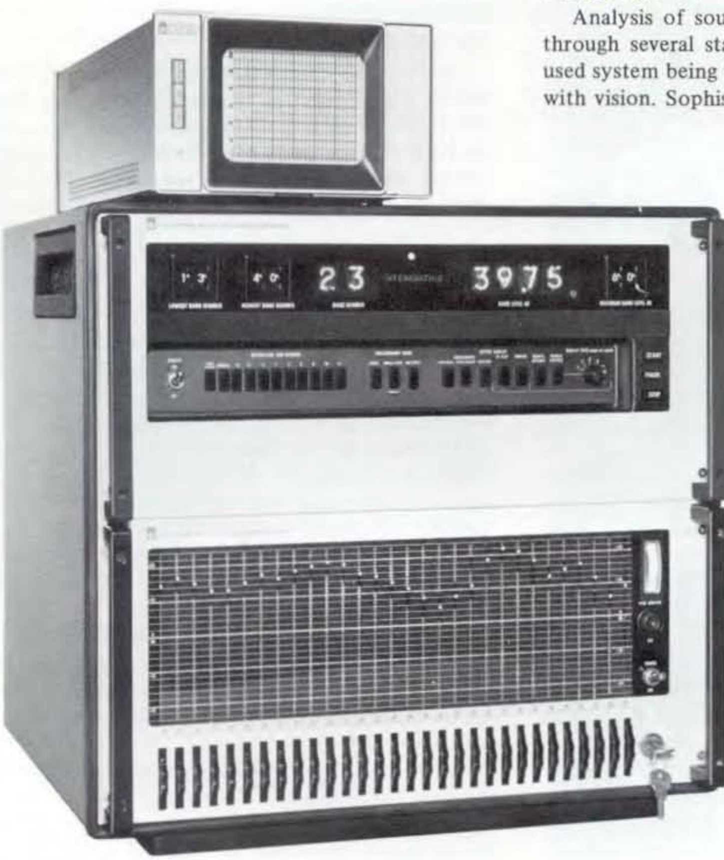
Long Time or Real Time?

This hurried world of ours substantiates the maxim that "Time is of the essence." The pressures placed upon design and development activities leave little doubt that solving engineering problems in real time is a capability most engineering activities must have, in order to remain virile and competitive. Consider the fact that the GR 1921 Real-Time Analyzer will measure an unknown noise spectrum and give you the corresponding Stevens loudness level, all in 0.5 second! Or that acoustic recordings of swiftly-moving vehicles, even planes, are accomplished so quickly that instructions for corrective measures can be transmitted to the vehicle, received and effected, and the results recorded while vehicle and station are still in a line of sight! This analyzer is a major contribution to active and meaningful research, development, and production.

Quickly summarizing, we can say that, when an analysis must be completed on-line, when large quantities of data are to be analyzed, or when a series of contiguous spectra is desired (e.g., automobile and airplane pass-by studies), real-time analysis is essential.

Background

Analysis of sound and vibration phenomena has passed through several stages, the most primitive and still widely used system being that of the human auditory senses coupled with vision. Sophisticated electronic systems in the second



Type 1921 Real-Time Analyzer, with accessory Type 1921-P1 Storage Display Unit.

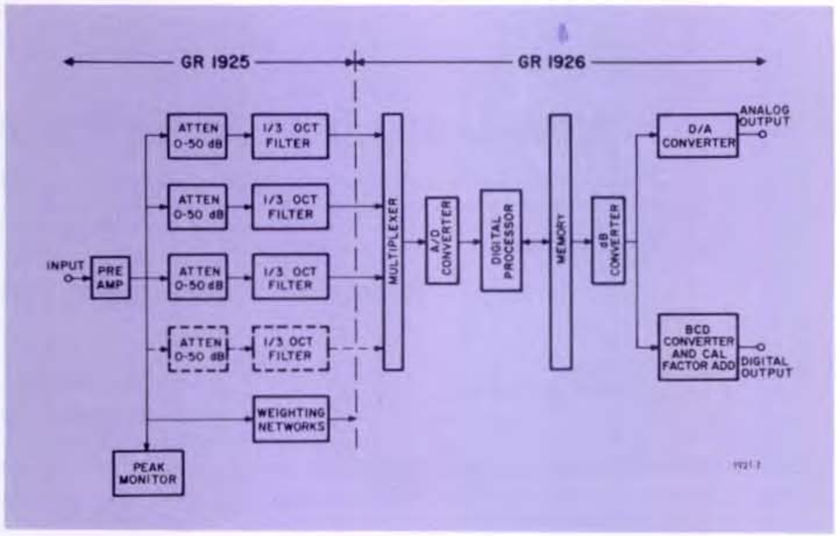


Figure 1. GR 1921 Block diagram.

stage helped to generate records of tests that were far more useful and durable. These systems, still employed, include sound-level meters, vibration meters, and swept (or serial) analyzers of many types. The advance from primary to second stage, unfortunately, did not provide a fundamental and necessary asset – the ability to analyze rapidly and automatically.

Looking back through the past three decades, we find an increased emphasis upon the vital need to cope with noise and vibration problems arising from technical advances in propulsion and other fields. Engineers have been forced to collect huge amounts of data, often laboriously, in order to analyze the complex problems involving various types of machinery, vehicles, and the disturbances that contribute so heavily to noise pollution of our living environment.

Some Design Views

An analysis of a time-varying signal, conducted with a conventional 1/3-octave serial analyzer, can employ a tape loop, which allows the test signal to be played back repeatedly as the analyzer is stepped to, or swept by, each center frequency. This method of analysis is quite time consuming and is not suitable for "on-line" operation.

Several methods of making real-time 1/3-octave analyses are possible. One technique employs conventional 1/3-octave analog filters followed by a set of analog detectors, or signal averagers, and a scanner. A dynamic range of no more than 30 to 40 dB is possible in an analog square-law detector (range changing in a real-time analyzer is not possible), and this is inadequate for many applications.

Digital filtering and the discrete Fourier Transform have been known for some time, and the science of digital signal processing is now rapidly advancing. The Fourier Transform is a natural basis for designing a spectrum analyzer; unfortunately, though it is easy to apply and to understand for a simple periodic signal, its use in analyzing complex or random signals is not fully understood and processors are only now available with sufficient speed to cover the audio range in real-time. Though a purely digital, real-time, 1/3-octave analyzer with wide-dynamic range is within the state of the art, it would be a very expensive device. So called recursive digital filtering must be ruled out because present-day processors are still too slow for full audio range 1/3-octave processing in real-time.

Our Design Choices

In the GR 1921 Real-Time Analyzer, we have chosen analog filtering and digital detection, taking advantage of both techniques to give us greatest accuracy and dynamic range. It is designed for sound and vibration work that often involves signals which may be random or totally unspecified. As a real-time analyzer it has many filter-detector channels, all energized concurrently by the signal to be analyzed. Our packaging permits independent use of the two basic units – the GR 1925 Multifilter[1] and the GR 1926 Multichannel RMS Detector. (See Figure 1.)

The GR 1925 Multifilter supplies summed and scanned outputs, in addition to its parallel output, providing for its use in a variety of spectrum-shaping, equalizing, and spectrum-generating applications. A signal connected to the GR 1925 drives all attenuator-filter channels simultaneously. Gain in each channel is calibrated and adjusted in 1-dB steps over a 50-dB range. The attenuators may be used to compensate for over-all measuring-system response errors. They also serve to de-emphasize, or pre-whiten, the input spectrum, thereby increasing the effective dynamic range of the GR 1921 system to as much as 95 dB.

One-third-octave filters, ranging from 3.15 Hz to 80 kHz and complying with current IEC and USA Standards,* are available in regular versions of the GR 1921. Octave-band filters ranging from 4 Hz to 16 kHz are also standard and

[1] Kundert, W. R., "A Calibrated Spectrum Synthesizer," *General Radio Experimenter*, October 1968. The GR 1925 may be purchased as a separate unit.

*IEC Publication 225-1966 and USA Standard USAS S1.11, 1966 Class 3 (High Attenuation).

available; narrow-band filters ranging from one-third to one-tenth octave are available on special order. The multi-filter chassis will accommodate up to 30 filters. Peak monitoring against overload is a standard feature, as well as mechanical key-locking of attenuator settings.

The GR 1926 detector can be used independently to measure as many as 45 signals simultaneously, which may be derived from multiple transducers or other sources. It can be considered as a 45-channel voltmeter with near-ideal characteristics. The detector operates by sampling each input channel and feeding these samples to the digital circuits that are time shared on all channels. It is available in 30- or 45-channel versions.

The detector simultaneously computes the rms level for each filter channel in the GR 1925. Up to 1024 samples are taken from each channel; the samples are converted to a digital binary number and squared. The squared values are accumulated in a memory register where spaces are provided for all channels.

Single integration or measurement periods are adjustable in nine octave steps ranging from 1/8 to 32 seconds. For integration periods of 1 second and longer, 1024 samples are taken from each channel during the integration period. For shorter integration periods the number of samples is proportionately reduced, with a minimum of 128 samples being taken in a 1/8-second integration period. At the end of an integration period, the sum-of-squares value is converted to decibels for output presentation. A single answer can be fed to a high-speed receiving device in about 15 μs, and all channel levels can be presented at the output in less than 1 ms.

Output data are presented simultaneously in digital (BCD format) and analog forms, and on a front-panel visual numeric display. Control signals and format are designed to permit output interfaces with a digital computer, printer, oscilloscope, automatic dc step-chart recorder, or X-Y plotter. A panel control allows the operator to add a scale factor to the digital output. The detector requires an input signal level of 1 volt rms for full scale maximum output. The output indication corresponding to a 1-volt input signal is adjustable from 60 to 159 dB in 1-dB steps.

Band numbers are presented with each output level and are set to correspond to the USA Standard Band Numbers for 1/3-octave filters (USAS S1.6 - 1967) for the particular version of the GR 1921. The number of channels to be displayed is controlled by the LOWEST BAND NUMBER and HIGHEST BAND NUMBER control settings.

Two extra internal channels are included, to calibrate zero level and full scale on the 1926. They can be measured during each integration period to monitor calibration continuously, if desired. The calibration channels are also intended to be used to set up analog output equipment, such as a reader or oscilloscope.

DIGITAL DETECTION ADVANTAGES

Wide Dynamic Range

There is really no fundamental limitation on dynamic range once a signal is converted to digital form, as there always is in the analog world. In the GR 1921 analyzer, only the multiplexer and A/D converter stretch analog techniques,

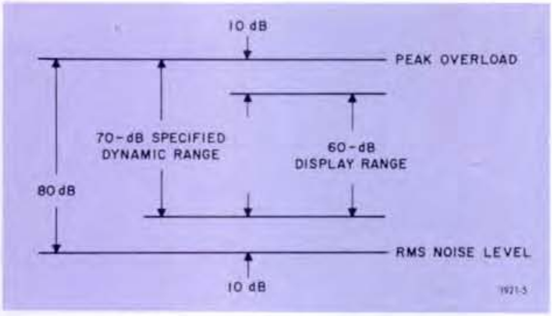


Figure 2. Dynamic range capability.

and these circuits have a dynamic range, measured from overload to noise level, of more than 80 dB. In the digital circuits it is only necessary to carry enough significant digits to attain the desired dynamic range.

The dynamic range situation for the GR 1921 analyzer is illustrated in Figure 2. On a per channel basis, the rms noise level is more than 80 dB below the peak-overload level. The specified dynamic range, the range of instantaneous levels for which signal-to-noise ratio is greater than 10 dB, is 70 dB. The display range extends from a level at least 10 dB above the noise floor to a level 10 dB below the peak-overload level.

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J. A. Lapointe holds degrees from Northeastern University (BSEE-1957 and MSEE-1959). At Raytheon Company he performed radar system circuit design, was responsible for systems design and project management at Sanders Associates, and acted as consultant at Signatron. In 1967 he joined GR, specializing in digital signal processing techniques. He is a member of IEEE and Eta Kappa Nu.

G. R. Partridge received his PhD(EE) degree from Yale in 1950 and taught at Purdue University during 1950-55 (Associate Professor EE). He joined Raytheon Company in pulse-communication work in 1955, transferring to GR in 1962 for design work in pulse generators and amplifiers. In 1958 he published ***Principles of Electronic Instruments*** (Prentice-Hall) and is author of numerous technical papers. His memberships include IEEE, Acoustical Society of America, Tau Beta Pi, Eta Kappa Nu, Sigma Xi, and he is a registered Professional Engineer in Massachusetts.

Figure 3. Example of analyzer/computer system.

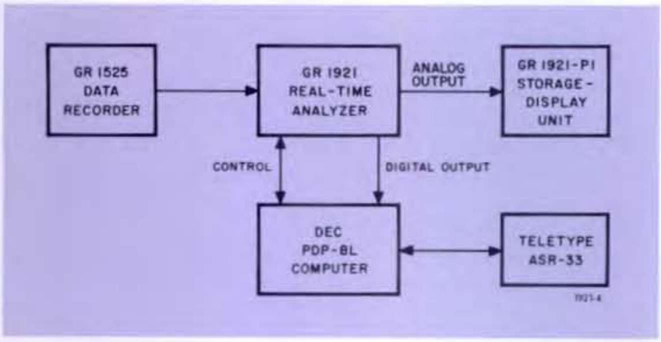


Figure 4. Block diagram of typical analyzer/computer system.

This ensures a crest-factor capacity, even at full scale, of 10 dB. Though instantaneous levels (samples) over a range of more than 80 dB are taken into account in computing band levels, answers in the lowest 10 dB are considered as noise and are therefore suppressed. Similarly, answers in the top 10 dB are displayed as overload, having encroached too closely upon the crest factor.

Computed band levels that exceed the upper limit of the display range are identified in the output by adding 800 dB to the erroneous digital signal with a resultant amplitude jitter in the corresponding analog level.

Accuracy

The digital approach used in the GR 1921 circumvents the accuracy limitations of analog methods. The computations can be performed to any desired resolution and, of course, all signals are processed with the same computer, so channel-to-channel uniformity is not a consideration.

True Integration Characteristics

The samples taken from each filter channel are squared and averaged to determine the band levels. When the "start" button is pushed, the detector samples the filter channels until the pre-selected integration time has elapsed. Samples are taken only during this period, and all samples are weighted equally in the result. The integration is thus linear, and the time period during which the measurement was made is accurately known. It is easy to vary the average rate at which samples are taken to produce any desired integration period. The GR 1921 has a choice of nine integration periods, ranging from 1/8 to 32 seconds.

By comparison, analog systems have several disadvantages. Long-time analog integrators in the quantity needed for a detector are expensive to the point of being impractical. To change integration time requires that a component in each detector channel be switched. This, too, is expensive, and electronically cumbersome. In an analog design, one must settle for a narrow choice of integration or averaging periods. Also, the only practical averaging method is simple RC smoothing. When presented with a signal, the output from this type of averaging circuit tends toward a finite saturation level. Events that occur early in the time signal are "forgotten" or leak off, and it is impossible to know exactly to what degree each event in the time signal has affected the answer. Several time constants are needed to bring this detector to saturation, the level for which it is calibrated, and this means that a longer time signal is needed for a given degree of statistical accuracy when the signal is random.

SYSTEMS AND SOFTWARE

The GR 1921 can produce data in prodigious quantities in a short time, posing the problem of data reduction. More important, it usually happens that the concern of the engineer is with a single number or possibly a suggested course of action, and not with the raw spectral information that is the product of the analysis process.

A small digital computer used in conjunction with the GR 1921 can perform any or all four basic functions:

- Analyzer Control - the computer can be programmed to select automatically all functions that are normally controlled manually by front-panel controls.
- Rapid Data Storage - the computer can ingest the results of a measurement in about one millisecond, thus freeing the analyzer to make another measurement. The stored results of the series of measurements are then typed out from the computer.

```
CLEAR

ID: WRK8
LOW BAND #: 14
HIGH BAND #: 43
INTEG TIME (SEC): 4

# MEASUREMENTS: 6
TIME BETWEEN MEAS (SEC): 2
OUTPUT: OCT, OCT MAX, TOCT MAX, SONES, PHONS, SONES MAX, SIL

START


WRK8

OCT BANDS
MEAS  15    18    21    24    27    30    33    36    39    42
   1  66.9 100.1 107.5 108.0 106.8 107.8 108.4 103.9  70.3  65.8
   2  64.4  99.4 103.9 107.4 105.9 104.4 102.5  96.5  62.8  65.5
   3  62.0  87.1  65.0  66.3  58.8  55.1  54.9  54.9  57.4  65.4
   4  61.6  96.5  94.5  89.1  62.4  55.1  54.9  54.9  57.8  65.5
   5  64.9 102.4 107.9 106.8  96.6  65.6  54.9  54.9  57.8  65.4
   6  96.3 108.4 106.4 105.3 105.1 103.4  93.0  62.6  58.0  65.9
 MAX  96.3 108.4 107.9 108.0 106.8 107.8 108.4 103.9  70.3  65.9

TOCT
BANDS
 MAX
  14  72.0  89.0  95.2  98.0 102.7 106.2 104.5 103.2 103.0 103.5
  24 103.2 102.5 100.5 102.7 102.2 102.0 103.2 103.2 102.2 105.7
  34 101.5  91.0 101.0 100.2  69.2  62.2  56.5  57.5  60.7  63.2

SONES MEAS
  1     2     3     4     5     6   MAX
 OVRLD OVRLD  24.9  44.6 116.7 191.5 OVRLD

PHONS MEAS
  1     2     3     4     5     6
 124.2 119.6  86.4  94.8 108.8 115.9

SIL   MEAS
  1     2     3     4     5     6
 107.6 104.2  56.2  57.4  72.3 100.4
                                                         1921-10
```

Figure 5. Data print-out.

- Comparison - the computer can compare the results of a measurement with a stored reference spectrum and make a decision depending upon their relation. For example, a noisy device can be rejected when its spectrum exceeds a preset limit. The computer also can be programmed to tell what limits have been exceeded, by what amount, and probable cause.
- Data reduction - the computer can be used to reduce the spectral information from the analyzer to obtain various acoustical ratings. For example, programs can be supplied to find loudness level, speech-interference level (SIL), and perceived-noise level (PNL).

Figures 3 and 4 show an analyzer/computer system and its corresponding block diagram. Data are fed to the analyzer from the GR 1525-A Data Recorder. The analog output from the analyzer drives the GR 1921-P1 Storage Display Unit, which displays each spectrum as it is transferred to the computer. The computer accepts digital data from the analyzer, and another link transmits control signals between the analyzer and computer. All communication between the system and the operator is via teletype.

This system and the operator work together in a "conversational" mode. The computer presents a series of questions and the operator replies; the questions concern desired frequency range, integration time, time between measurements, and quantity of measurements. After the operator has supplied this set-up information, the computer asks which of a series of standard data-reduction operations it should perform. This information is supplied by the operator and then, at the touch of a key, the system performs the required measurements and types out answers in a standard format. The system is capable of making a rapid series of 1/3-octave measurements with selectable time delay between measurements. It computes octave band levels from 1/3-octave levels, and it determines the maxima from corresponding band levels in the series of spectra using either octave or 1/3-octave data. It can also define maximum-loudness and maximum-speech-interference levels.

An example of the typed-information output from the system is shown in Figure 5. The underlined data are instructions supplied by the operator; all other data are computer-controlled output.

Basic software packages are available to compute octave-band levels, SIL, PNL, Stevens loudness level, and, if desired, the maximum of these in a sequence of different measurement intervals. Software packages to control the analyzer have been developed for use in custom systems. Other programs can be made available for custom analyzer/computer systems.

APPLICATIONS

The one-third-octave analyzer is the heart of most acoustic noise-analyzer systems. This relatively broad, constant-fractional-bandwidth method of analysis is, for a variety of reasons, well suited to airborne noise analysis and control work.

On-line operation is made practical by the high speed and accuracy of the GR 1921. As mentioned in the introduction, measurements of aircraft fly-by and automobile pass-by are so rapidly assimilated that corrective measures can be accomplished during the test period. This approach, which is not possible using a tape-loop serial system, can greatly reduce the cost of testing.

Time-varying signals are handled by the analyzer so fast as to permit liberal application of spectrum analysis. It is possible to produce three-dimensional plots, with suitable display devices, to present spectrum versus time. Or, a computer system might be programmed to present long-time average spectra in the section of a time-varying signal, where the changes are not significant, while retaining the fine detail in other areas.

Volume-data-reduction work is economically accomplished through intelligent use of an analyzer/computer system. This is especially true when compared to the conventional method of employing a swept analyzer and recorder and reducing data by means of tables and a desk calculator. A comparison table, showing relative times to measure a spectrum and then to compute the corresponding Stevens loudness level, is shown below.

Conventional Means		1921/PDP-8L System	
Time for analysis	20 s	Time for analysis and computation of loudness level	0.5 s
Read chart levels and look up loudness indices	300 s		
Calculate loudness level	180 s		
Total	500 s	Total	0.5 s

The saving in time is clearly evident; one year's work by conventional methods is now accomplished in a few hours! Even when a general-purpose computer is used to find loudness level from spectral information, considerable time is required to prepare data for insertion in the computer. In addition, waiting for availability of a general-purpose computer greatly reduces the attractiveness of this approach.

Production-line product testing has seldom employed spectrum analysis because of the long time required to obtain accurate measurements. The real-time analyzer eliminates this objection and potentially makes it a powerful production-line tool for fault diagnosis. The use of a small computer in a production-line system can increase effectiveness even further. The computer could be programmed to compare measured spectra with reference spectra, suggesting possible faults and repairs to be made to the rejected product.

Transmission-loss measurements, in which dynamic range and signal-to-noise-ratio may be critical, are simplified through use of the calibrated adjustable channel attenuators on the GR 1925 Multifilter. A block diagram of a system for transmission-loss testing is shown in Figure 6. The GR 1382 Random-Noise Generator and the multifilter provide a source-side spectrum, shaped to provide a uniform signal-to-noise ratio versus frequency on the receiving side. The multifilter on the receiving side is set to have a transmission pattern that is approximately the inverse of the transmission pattern on the sending-side multifilter, taking into account frequency-response errors of the transducers. This results in a direct-output plot of transmission on the recorder, while minimizing the source-power requirements and maximizing tolerance to interfering signals on the receiving side.

Radiated power measurements can be made by dividing the same integration period among a specific number of microphones. A switch closure to ground, at the appropriate control line on the GR 1926 Detector, divides the selected integration period into 2, 4, 8, or 16 segments, corresponding to the number of microphones. Alternatively, the position of a moving microphone can be synchronized with the measurement interval in order to obtain a space integration.

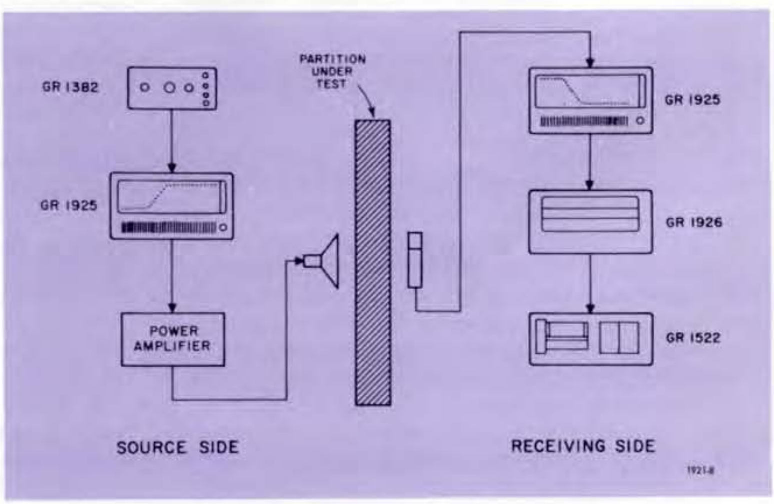


Figure 6. Block diagram of transmission-loss measurements.

AFFINITY FOR ACCESSORIES

The GR 1921 analyzer has great flexibility in operation with accessory equipment.

It can be expanded to 45 channels by use of the 45-channel version of the GR 1926 detector and two GR 1925 multifilter units.

It supplies power to and interfaces with the GR 1560-P40 Preamplifier, suitable for high-impedance transducer inputs.

It operates with the GR 1525 Data Recorder and similar tape recorder units that have good dynamic and frequency ranges and stable characteristics. Of particular interest is use of the maximum-level dB control on the analyzer to restore the calibration factor lost in the recording process. This control is used to make the analyzer read out in decibels, with the reference level the level of the original signal.

It is compatible with the new GR 1522 DC Recorder* (bottom unit, Figure 7a), an analog recorder much faster than conventional X-Y plotters. The new recorder operates synchronously with the GR 1921. A short dwell period as each band level is selected allows the recorder pen to settle, producing a neat bar graph with standard scale factor. The analyzer/recorder combination and a sample chart record are shown in Figure 7.

It can supply output data at rates up to 360 band levels per second. The MDS Series 800 Hi-Speed Digital Printer is available on special order from General Radio. The printer should be used when most accurate numerical data are required. Note that the GR 1921 is capable of driving a printer and analog recorder simultaneously.

*Refer to page 16.

Figure 7a. Visual/record assembly including, from top to bottom, GR 1921-P1 Storage Display, GR 1926 Multichannel RMS Detector, GR 1925 Multifilter, and GR 1522 DC Recorder.

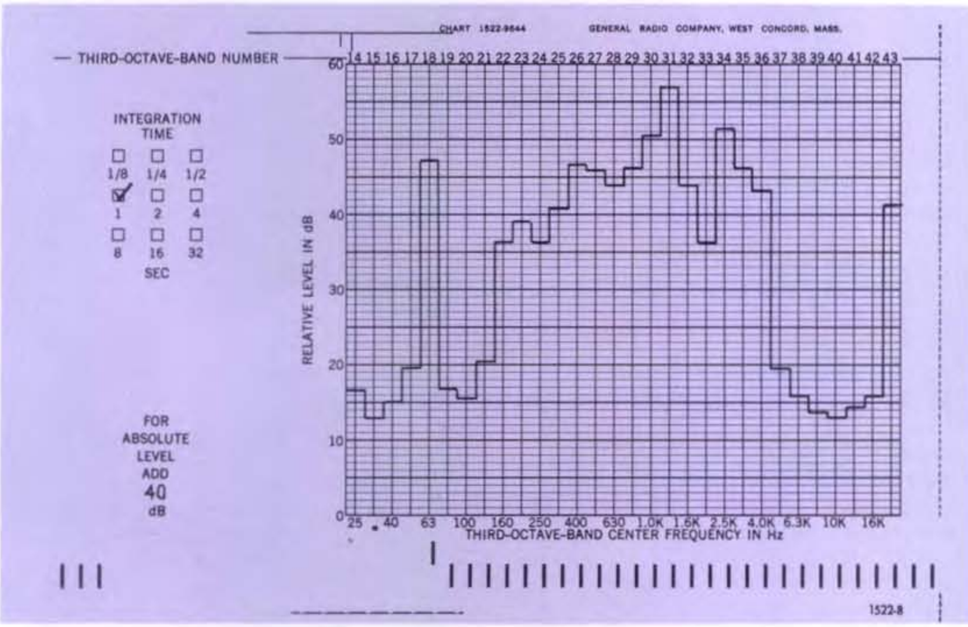


Figure 7b. Typical GR 1522 graphic recording.

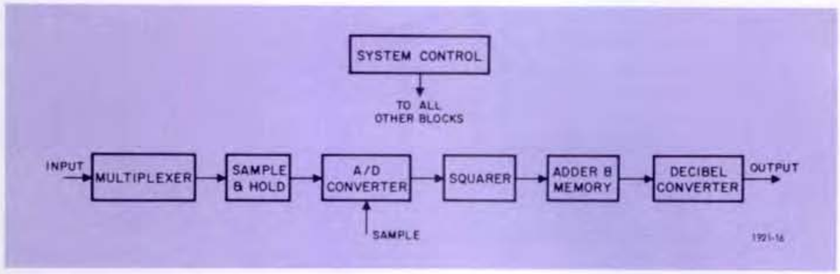


Figure 8. Block diagram of digital detector.

The Houston Instrument Series 6400 (with 024 option) Omnigraphic Recorders are available on special order from General Radio when an X-Y recording format is desired. These recorders use fan-fold charts that load automatically, thereby taking advantage of the speed of the 1921 units. Conventional X-Y recorders are not recommended because of slow loading time.

The GR 1791 Card Punch Coupler is available to couple the GR 1921 units to an IBM 526 Printing Summary Punch, permitting transfer of band numbers and levels to standard IBM cards.

The GR 1921-P1 Storage Display Unit (top unit, Figure 7a) will provide a rapid visual display of spectra from the analyzer. It is a slightly modified version of the Tektronix Type 601 Storage Display Unit. Three CRT display modes can be selected by the GR 1921 – NON-STORE, STORE-ERASE, and STORE-NO ERASE. When any of the three scope modes is selected at the GR 1921, data are fed out at a rate of one band per millisecond. When X-Y PLOT or DATA PRINTER output modes are selected, the scope remains in operation but is fed data at the rate set by the DISPLAY RATE control. In this manner, the unit continues to monitor output when recording devices are used.

A CLOSER LOOK AT THE DETECTOR

A unique design deserves some elaboration in order to stress its qualities and advantages. This section expands upon our previous presentation of design details and refers to Figure 8.

The *Multiplexer* is an electronic switch with up to 45 inputs and one output. The input channels are switched on in sequence by command from the system control, using addresses coded in an 8 X 6 matrix. Three extra input channels are available; one is left unused, the other two are reserved for calibration purposes. Calibration of zero level is achieved by addressing a channel with grounded input; the second calibrate channel is supplied with a precision dc voltage to produce a full-scale signal level of 60 dB for this channel.

The *Sample-Hold Circuit* takes two samples of its input simultaneously. One microsecond after the samples are taken, a decision is made to select the one sample that is within the linear operating range. This decision, designated "coarse-range," is also used by the A/D converter and the squarer. Polarity inversion is switched in, if needed, so that a positive voltage is always delivered to subsequent circuits.

The A/D Converter uses floating-point binary arithmetic to achieve a constant-percentage accuracy. After the coarse-range decision (as noted above), fine ranging is performed to

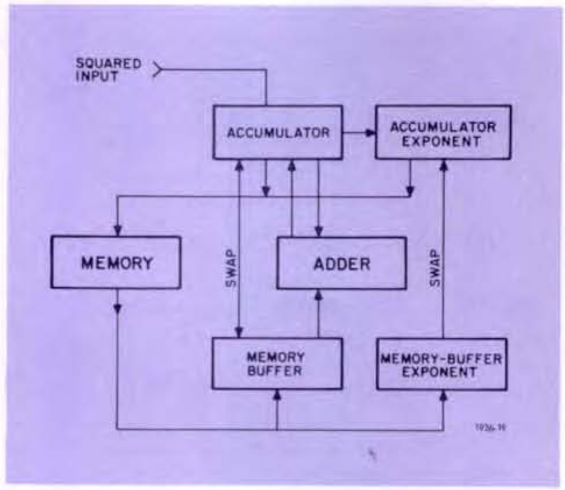


Figure 9. Block diagram of computational mode.

locate the signal in a 2:1 amplitude group. Next a 3-bit group of digits completes the process, for a total effective conversion range of 15 bits.

The *Squarer* accepts binary inputs from the sample-hold circuit and the A/D converter to generate the squared value of each input sample. Its output is in a floating-point format that is consistent with the arithmetic of the computer.

The computational mode is dominated by the *ADDER AND MEMORY* block of Figure 8, shown in more detail in Figure 9. This represents a very simple, but true, digital computer. The squared numbers are delivered to the accumulator in two parts: a mantissa and an exponent. Simultaneously, the number in memory is taken into the memory buffer as a mantissa and an exponent; the two exponents are then compared. Since only the accumulator has provision for shifting digits, the number with the smaller exponent must be in the accumulator. If a swap is necessary to get the number with the smaller exponent into the accumulator, a logic circuit orders the swap made. Next the exponents are made equal. The accumulator bits are shifted to the right by one place and the accumulator exponent is increased in value by one digit.

Once the exponents in the accumulator and memory buffer are equal, the digits in corresponding stages of the mantissas have the same weight and may then be added.

The output of the adder is returned to the accumulator for one last check. If the result of the addition involves a "carry",

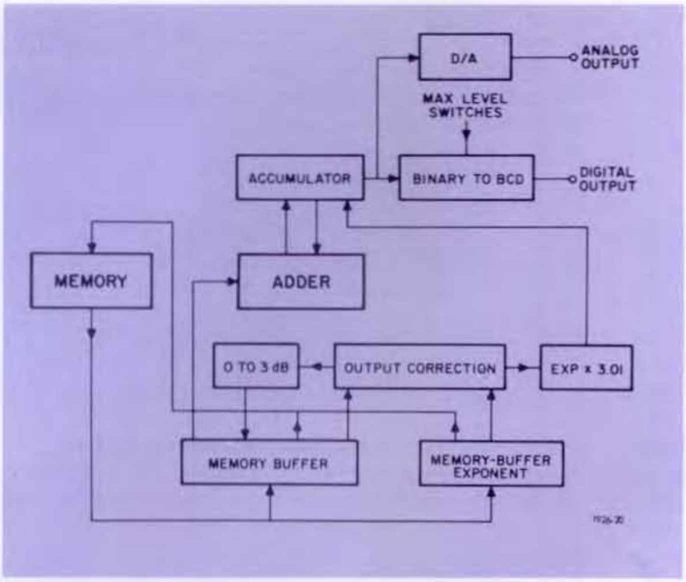


Figure 10. Block diagram of output mode.

so that the accumulator contains 16 rather than 15 bits, then the mantissa must be cut back to size. This is done simply by shifting the mantissa one place to the right and increasing the exponent again by one digit. Therefore, if there is a carry to 16 places, one more shift and add will be needed. If there is no carry, the accumulator is finished with its work on this particular squared-input number.

The contents of the accumulator and accumulator exponent are put back into the memory unit. Another squared number is presented to the accumulator, and the whole process repeats.

The output mode begins when all data gathering is completed. Figure 10 shows the sequence of data flow. In the first step, the stored sum-of-the-squares is drawn out of the memory into the memory buffer. This number is fed to a block marked *OUTPUT CORRECTION*. If the time of integration was less than 1 second, the sum of squares is a smaller number than if the time had been 1 second or longer (as the time is reduced below 1 second, fewer samples are taken). The *OUTPUT CORRECTION* block multiplies the sum of the squares by 2, 4, or 8 if the integration time is 1/2, 1/4, or 1/8 second, respectively. The numbers derived from the mantissa in the sum of squares give a contribution of between 0 and 3 dB to the final answer. This contribution is determined in the *0 TO 3 dB* block of Figure 10. The numbers derived from the exponent in the sum of the squares must be multiplied by 3.01.

The circuits in the *MEMORY-BUFFER* and the *MEMORY-BUFFER EXPONENT* blocks are clocked JK flip-flops. They can accept new data into storage as the previously held data are read out. The next step in the output sequence takes advantage of this fact; the sum of the squares, in binary format, is returned to the memory for future use at the same moment that the result of the *0 TO 3 dB* determination is fed into the memory buffer. Simultaneously, the product of *EXP X 3.01* is deposited in the accumulator. The accumulator and memory buffer are then connected to the adder, which gives the sum of the *0 TO 3 dB* and the *EXP X 3.01* contributions. The adder output is the binary representation, in decibels, of the sum of the squares that was just taken from, and returned to, memory.* The adder output is then transferred to the accumulator, which now contains the decibels in binary form. This result is converted to a binary-coded decimal (BCD) in 1-2-4-8 format. If the front-panel MAXIMUM BAND LEVEL dB control calls for the range of answers to be other than 0 to 60 dB (e.g., 20 to 80 dB), an appropriate constant scaling factor in decibels is added at this time.

The final output consists of

(1) BCD representation of the decibels calculated from the sum of the squares plus the calibration factor, if any, added by the MAXIMUM BAND LEVEL dB control, and

(2) An analog voltage proportional to the number of decibels, *not including* the calibration factor.

If the exponent in the sum of the squares is too large, the instrument concludes that an overload condition existed in the input signal. It then causes the answer to be reported as 800-some or 900-some dB's. Answers up to 62.75 dB (not counting any additional decibels added by the MAXIMUM BAND LEVEL dB control) are reported correctly.

*Note that the square root is never taken explicitly; the conversion is directly from the sum of the squares to decibels.

Acknowledgment

The authors gratefully acknowledge the support of a number of General Radio engineers who have contributed greatly to this development program. Jim Esselstyn was responsible for the mechanical design and packaging of the entire GR 1921 system. Matt Fichtenbaum participated in early work on the detector and might properly be described as the computer section architect. Dave Nixon was involved in the early cost and feasibility study on the detector. Ralph Anderson designed all the power supplies for the system. Jim Faran designed the output D/A converter for the detector and did the original work on the input A/D converter. Carl Woodward has been involved in interfacing accessories with the system. Arnold Peterson acted as consultant during the entire program.

Complete specifications for the GR 1921 are included with this issue as a tear sheet, removable for insertion in GR Catalog T.

Some Notes on Digital Detection

A digital detector computes rms voltage, starting with the sample variance formula

$$\sigma^2 = \frac{1}{n} \sum_{i=1}^{n} V_i^2 \qquad (1)$$

where n is the number of samples included in a measurement and V_i are the input voltages that occur at discrete instants of time. To obtain an output in decibels, it is necessary to find

$$\mathrm{dB} = 10 \log_{10} \left(\frac{\sigma}{\sigma_0}\right)^2 \qquad (2)$$

where σ_0 is the desired reference level.

The input signal to be measured is a continuous function of time. It is sampled only at discrete instants of time when it is accepted for digital processing. Two questions naturally arise. How many samples are needed for a measurement, and when should they be taken? To answer these questions, we must examine both random and periodic input signals.

Random Inputs

First, consider the case of a stationary random input signal with a zero mean voltage and a Gaussian amplitude distribution.

Curves of measurement repeatability versus independent signal samples are shown in Figure 11. These curves are used in pairs and represent the upper and lower limits of expected measurement variations. They are commonly known as confidence limit curves, where the confidence in percent specifies the probability that a measurement lies within the area bounded by the curves. The most efficient sampling scheme for purely random inputs is uniform sampling, with a sample spacing equal to the correlation time of the input. Samples must be statistically independent to be effective for measurement purposes and, hence, must not be taken at intervals closer than the correlation time.

For example, a white noise limited to the frequency range of 0 to BW hertz has a correlation time of $\frac{1}{2BW}$ second This corresponds to the Nyquist rate, which is the required sampling rate for signal reconstruction. A lower sampling rate is perfectly acceptable for variance measurement.

Now consider the measurement of Gaussian noise with the GR 1921 analyzer. The GR 1925 filter bands have a constant fractional bandwidth and hence the correlation time will vary with the center frequency. For a 1-second integration time, 1024 independent samples will not be available from all bands. The number of independent samples available from a particular band will vary depending on the integration period selected. These factors were included, along with the results of Figure 11, to generate the composite confidence limits for 1/3-octave-filter center frequency and are shown in Figure 12.

Periodic Inputs

For stationary random inputs, we simply require enough measurement samples to obtain statistical stability (repeatability). With periodic inputs, however, we shall see that the sample rate is also a critical parameter. Consider the case where the input $V(t)$ is a sinusoid of unit variance, unknown frequency f, and phase ϕ.

$$V(t) = \sqrt{2} \cos (2\pi ft + \phi) \qquad (3)$$

With n samples of the input and a uniform sample spacing of δ seconds, the variance estimate is

$$\sigma^2 = \frac{2}{n} \sum_{i=1}^{n} \cos^2 (2\pi f i\delta + \phi) \qquad (4)$$

Errors in the variance estimate are primarily due to the number of samples n and the sample spacing δ. Phase angle ϕ can be regarded as a random variable that is uniformly likely to be

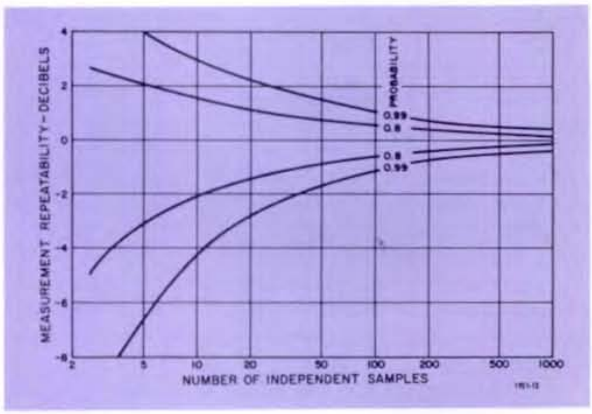


Figure 11. Confidence limits of measurements (in general).

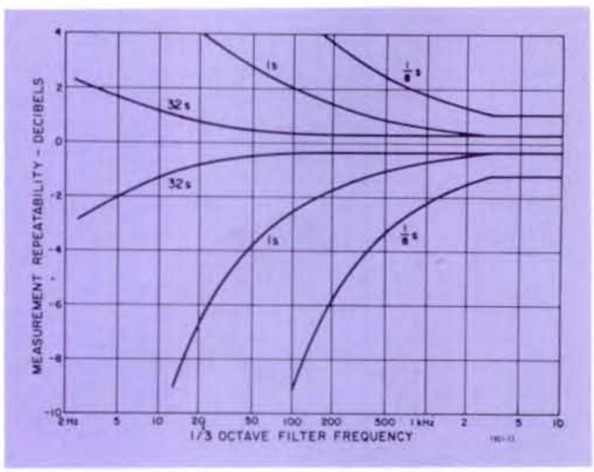


Figure 12. Composite 99% confidence limits of measurements (1/3 octave).

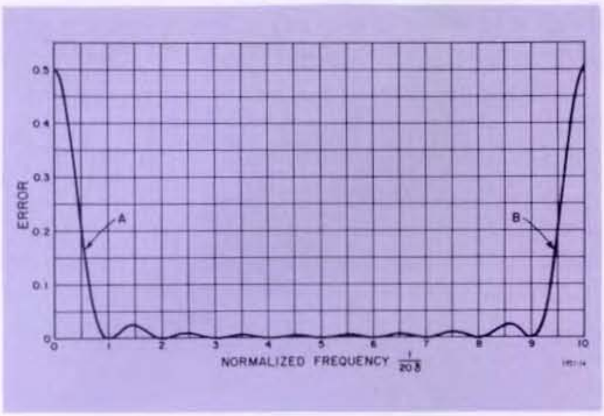


Figure 13. Measurement errors expected for fixed sampling rate.

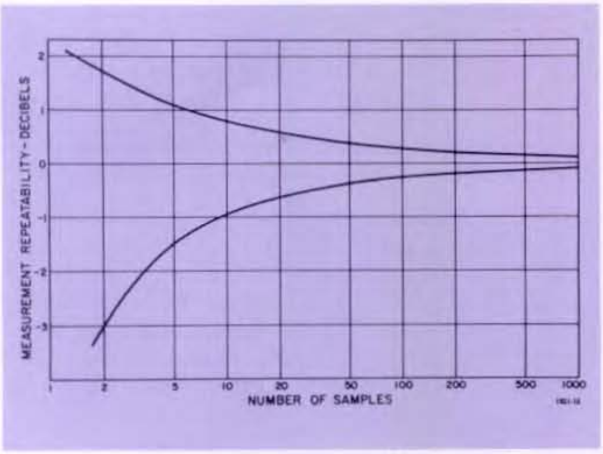


Figure 14. Range of repeatability of measurements.

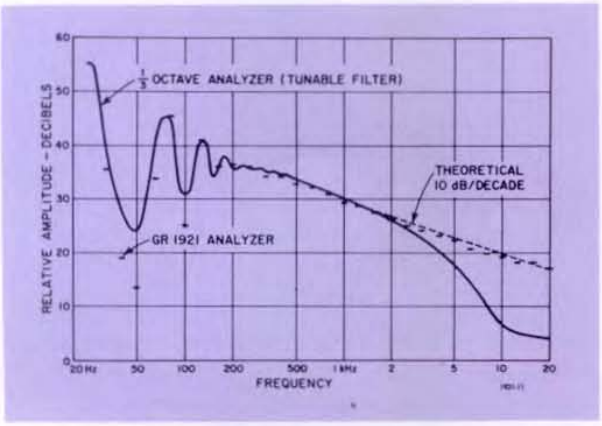


Figure 15. Measurements comparison at 25 Hz -- Tunable 1/3-octave filter versus GR 1921.

between 0 and 360 degrees. Averaging the effects of ϕ in equation (4) and subtracting the true variance, we have the mean-squared error result:

$$\epsilon^2 = \frac{1}{2}\left(\frac{\sin 2\pi n f}{n \sin 2\pi f}\right)^2 \qquad (5)$$

Equation (5) is plotted in Figure 13 for the case of $n = 10$, to show the effects of input frequency f relative to sample spacing δ. The largest errors occur in the frequency ranges A and B. The remaining error lobes are relatively small in magnitude.

The large error lobes repeat at input frequencies that are multiples of one-half the sample rate. Because these frequencies lie within the input-frequency range of the GR 1926 detector, uniform sampling is unsatisfactory.

Modulating the sampling rate will cause the frequencies to shift. A shift of the sampling frequency during a measurement interval will shift the frequencies at which peak errors occur, thus distributing the error. This principle is incorporated in the 1926 detector design. The sample rate is swept during each measurement period, over a 2:1 frequency range. With 1024 samples for measurement, the worst-case error is reduced to about 0.2 dB.

The measurement repeatability for various values of n, assuming a modulated sampling rate, can be approximated by integrating curves like that of Figure 13. The results are plotted in Figure 14, which shows the dependence of repeatability upon n, the number of samples per measurement. These curves represent the 1σ limits of the range of measurement repeatability.

So far we have considered only the special case of a sinusoidal input signal. Other periodic signals should also be considered. A gated sinusoid (tone burst), for example, is measured with less than 1024 samples. The sample number reduction is in proportion to the duty factor. The measurement repeatability is determined by the number of effective samples and is specified by the curve of Figure 14.

A Difficult Spectrum

Consider the output of a 1/3-octave filter set with a square-wave input. At low frequencies there is only one harmonic in the the output of each 1/3-octave filter. At high frequencies, however, the filters' outputs are periodic sequences consisting of the step-function responses of the filters. The results of analyzing a 25-Hz square wave with a tunable 1/3-octave filter and analog detector are compared with the results of measuring this same signal with a GR 1921 Real-Time Analyzer in Figure 15. A dashed line with a slope of 10 dB/decade shows the theoretical spectrum shape.

At low frequencies there is slight disagreement because of a difference in filter shapes. At a frequency of 10 kHz, a measurement error of about 13 dB is made by the analog detector because of the high crest factor of the filter output signal, while the expected measurement repeatability from the GR 1921 (and indeed the measured variation) is less than 1 dB.

—J. A. Lapointe

Rayleigh-Distributed Noise

A Method for Reproducing a Singular Type of Interfering Background Noise

Noise that appears at the output of envelope detectors such as those in some radio or radar systems typically has the Rayleigh amplitude distribution rather than the Gaussian. The Gaussian density distribution function (Figure 1) is the familiar symmetrical bell-shaped curve; the Rayleigh distribution function (Figure 2) is quite nonsymmetrical, being zero for negative values of noise voltage. In experiments on signal detection or intelligibility, it may be important to reproduce accurately the type of interfering background noise. Because the Rayleigh distribution function is very small where the noise voltage is near zero, it cannot be approximated accurately by full-wave rectification of Gaussian noise.

One of the most common examples of a voltage having Rayleigh distribution is that seen at the output of a detector whose input is narrow-band random noise. Generating Rayleigh noise in this manner involves amplification, filtering, detecting, and smoothing. This necessarily results in considerable reduction of the bandwidth of the Rayleigh noise compared to that of the original Gaussian noise.

There is a procedure for starting with Gaussian noise and generating Rayleigh noise which does not require filtering and smoothing and the attendant bandwidth reduction. Start with two statistically-independent Gaussian random noises, square each and add, and take the square root of the sum. These analog operations are indicated in Figure 3. Proof that this procedure yields the Rayleigh distribution is given below. The argument is similar to that used in solving the famous problem of the random walk in two dimensions (References 1, 2). The joint probability distribution function of two Gaussian random variables x and y, having rms values of 1 (that is, $\sigma = 1$), is

$$p(x, y)\, dx\, dy = \frac{1}{2\pi} e^{-\frac{x^2+y^2}{2}}\, dx\, dy.$$

Consider the distribution of the variable r where $r^2 = x^2 + y^2$. In terms of r,

$$p(r)\, r\, dr\, d\theta = p(x, y)\, dx\, dy = \frac{1}{2\pi} e^{-\frac{r^2}{2}}\, r\, dr\, d\theta$$

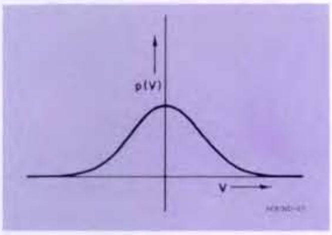
(where the element of integration $dx\, dy$ has been replaced by its counterpart in polar coordinates, $r\, dr\, d\theta$). Integrating over all angles,

$$p(r)\, dr = \frac{1}{2\pi} e^{-\frac{r^2}{2}}\, r\, dr \int_0^{2\pi} d\theta$$

$$= r e^{-\frac{r^2}{2}}\, dr, \qquad r > 0. \qquad (1)$$

This is the Rayleigh distribution function. The procedure for generating r from x and y is precisely that shown in Figure 3.

Figure 1. The Gaussian probability density distribution function.



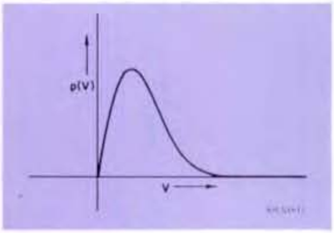


Figure 2. The Rayleigh probability density distribution function.

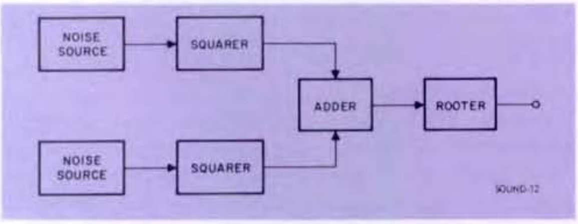


Figure 3. System for producing noise having a true Rayleigh distribution.

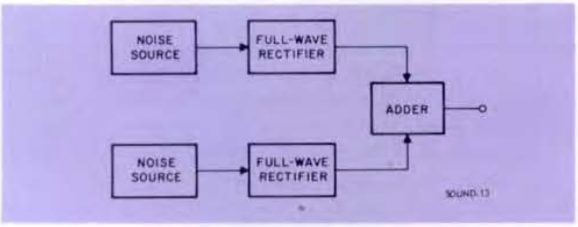


Figure 4. Simpler system for producing noise having approximately the Rayleigh distribution.

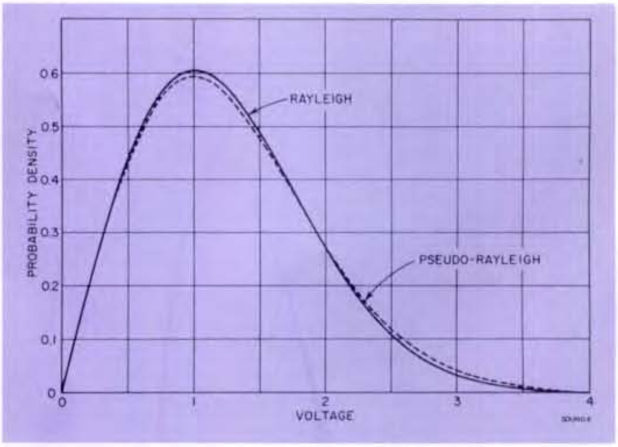


Figure 5. Comparison of amplitude distribution of pseudo-Rayleigh noise (suitably adjusted in amplitude) with that of true Rayleigh noise.

There is a much simpler procedure for generating a close approximation to the Rayleigh distribution. Shown in Figure 4, it avoids the use of the squaring circuits (squarer) and the square-root circuit (rooter).

To determine the actual amplitude density distribution function of the output of this system (pseudo-Rayleigh noise), we start with the Gaussian distribution function for each noise source,

$$p(x)\,dx = \frac{1}{\sqrt{2\pi}}\, e^{-\frac{x^2}{2}}\, dx,$$

where again, for simplicity, σ has been taken equal to 1. After full-wave rectification (mathematically, taking the absolute value) the distribution becomes, for each noise source,

$$p_1(x)\,dx = \sqrt{\frac{2}{\pi}}\, e^{-\frac{x^2}{2}}\, dx, \qquad 0 \leqslant x.$$

The probability that the sum $(x+y)$ will have a value between x_0 and $(x_0 + dx_0)$ is

$$p_A(x_0)\,dx_0 = \int_0^{x_0} p_1(x)\, p_1(x_0 - x)\, dx \cdot dx_0,$$

where the integration is taken over all possible values of x which could lead to a sum x_0. (Remember that $x > 0$ and $(x_0 - x) > 0$.) Substituting for $p_1 x$, and dropping dx_0,

$$p_A(x_0) = \frac{2}{\pi}\int_0^{x_0} e^{-\frac{x^2}{2}}\, e^{-\frac{(x_0 - x)^2}{2}}\, dx$$

$$= \frac{2}{\pi}\, e^{-\frac{x_0^2}{4}} \int_0^{x_0} e^{-\left(x - \frac{x_0}{2}\right)^2} dx.$$

Now, letting $y = x - \frac{x_0}{2}$,

$$p_A(x_0) = \frac{2}{\pi}\, e^{-\frac{x_0^2}{4}} \int_0^{\frac{x_0}{2}} e^{-y^2}\, dy$$

$$= \left(\frac{2}{\sqrt{\pi}}\, e^{-\frac{x_0^2}{4}}\right)\left(\frac{2}{\sqrt{\pi}} \int_0^{\frac{x_0}{2}} e^{-y^2}\, dy\right) \qquad (2)$$

$$= H'\!\left(\frac{x_0}{2}\right) H\!\left(\frac{x_0}{2}\right),$$

where H' and H are precisely the functions tabulated in Reference 3. When the values of x_0 are multiplied by 0.8 and the values of $p_A(x_0)$ are multiplied by 1.25 we find the agreement with the true Rayleigh distribution shown in Figure 5. These multiplications simply amount to changing the amplitude of the pseudo-Rayleigh noise; the area under the probability curve remains equal to unity. In the figure $p_A(x_0)$ is compared with the function $p(r)$ from Equation 1. The pseudo-Rayleigh noise has an amplitude distribution which is extremely close in form to the true Rayleigh distribution.

Examination of Equation 2 shows that the behavior of $p_A(x_0)$ when x_0 is very small is the same as for the Rayleigh distribution; however, the approximation behaves as $\exp\left(-\frac{x_0^2}{4}\right)$ instead of $x_0 \exp\left(-\frac{x_0^2}{2}\right)$ for large values of x_0. The latter behavior hardly shows on the linear plot for Figure 5. This difference does result in a detectable change in the number of peaks above a given level.

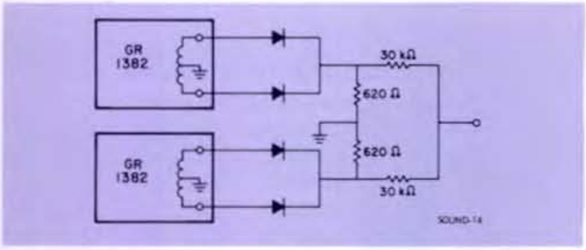


Figure 6. System for producing pseudo-Rayleigh noise.

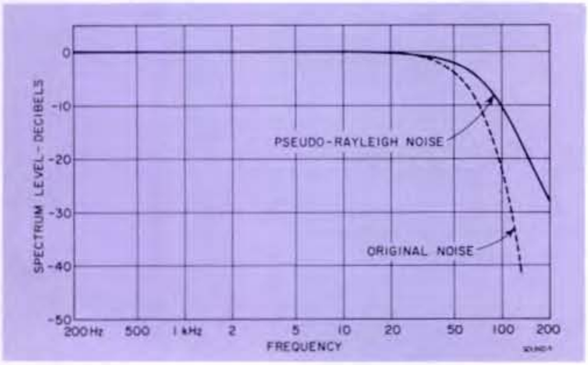


Figure 7. Comparison of pseudo-Rayleigh noise spectrum with the output spectrum of the GR 1382 generator.

A schematic diagram of how pseudo-Rayleigh noise can be produced from two GR 1382 Random-Noise Generators[1] is shown in Figure 6. High-conductance germanium diodes, such as 1N455 or 1N695, should be used for low forward voltage drop. A dc amplifier will be needed if it is necessary to reduce the output impedance level or to provide more power.

The effect on the spectrum of this method is shown in Figure 7. This is a comparison of the measured input and output spectra of the system shown in Figure 6. In Figure 7, the spectra are normalized to the same level at low frequencies. This method of approximating Rayleigh noise simply spreads the spectrum a little towards higher frequencies. Similar spreading probably will occur with the true Rayleigh system shown in Figure 3.

This easily assembled system for generating pseudo-Rayleigh noise includes only noise generators, diodes, and resistors. It produces noise having a close approximation to the Rayleigh amplitude distribution.

—J. J. Faran

A brief biography of Dr. Faran appeared in the March/April, 1969 issue of the *Experimenter*.

[1] Faran, J. J., "Random-Noise Generators," *General Radio Experimenter*, January 1968.

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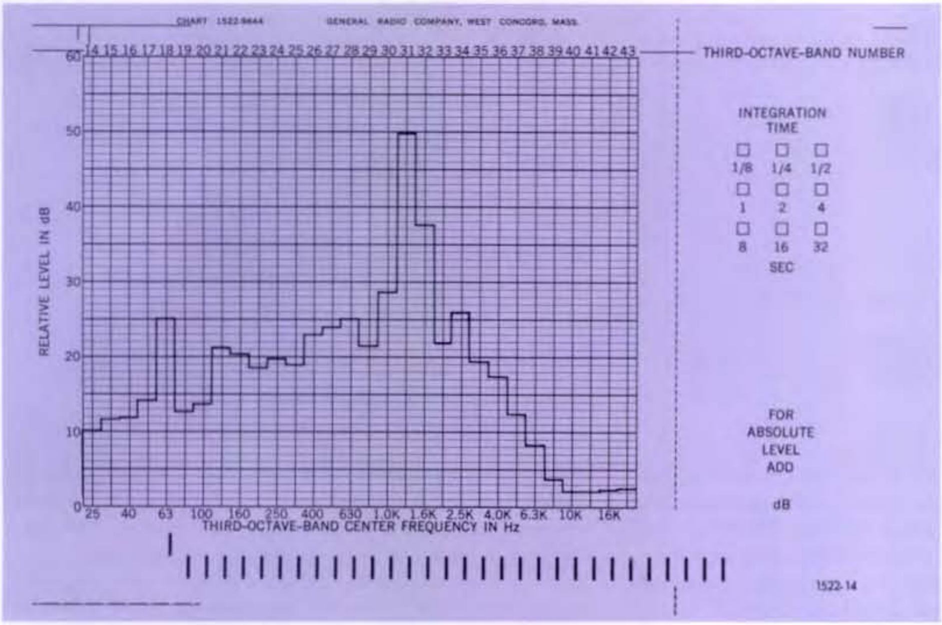

Figure 1. Typical recording programmed by control marks.

excursions beyond preset tolerances, and of time relationships between charts made on several recorders simultaneously.

- Reinforced mounting sheets (8½″ by 11″) are available for GR 1522 charts. Two adhesive strips are provided on the sheets to facilitate the mounting of recorded charts, which can then be filed in data notebooks.
- The Y-axis recording action of the GR 1522 is much faster than that of the X-Y recorder-65 inches/second as compared with 20 inches/second for the X-Y recorder.
- The X-axis of the GR 1522 is expandable manually or automatically, with local or external programming; it can range from 10 to 100 inches to permit detailed physical and temporal waveform investigation. The X-Y recorder is limited to 10 to 20 inches.
- Control of the X-axis motion is by digital rather than analog signals. For this reason it is possible to convert angle or shaft position to chart position directly, using an optical or commutator encoder.
- Isolated grounding provisions avoid ground loops between the GR 1522 and other instruments.
- External sync signals are available to operate several recorders synchronously.
- The GR 1522 is capable of use with small computers, provided proper interfacing is supplied.

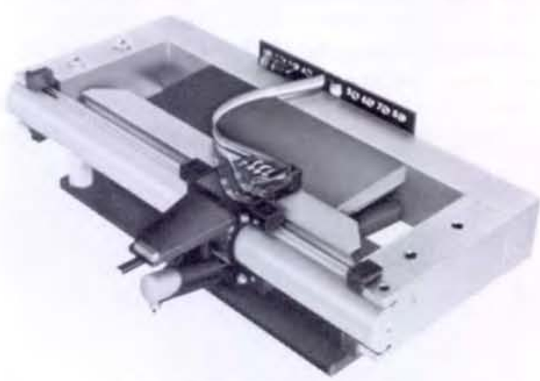

Figure 2. Servo-motor pen drive.

Some Design Facts

Backbone of the 1522 is a field-proven dependable servo motor,* Figure 2, consisting of an aluminum-wire coil wound on a light-weight plastic form. This coil assembly is supported by four miniature ball bearings gliding in a V-groove formed in a longitudinal iron bar. The iron bar is part of the magnetic path produced by a large permanent magnet that maintains a uniform magnetic field in the gap through which the coil assembly moves. When a current is applied through the coil, a force is generated perpendicular to the current path and magnetic field. This force moves the coil assembly to right or left, depending on the polarity of the current, imparting a true linear motion. The ball bearings cause very little backlash and have very little friction, even though the force is off center. (Force is applied in the magnetic gap, not in line with the bearings.) The pen and potentiometer wiper arm are mounted directly on the motor coil; this technique eliminates backlash between pen and wiper arm of the feedback potentiometer.

The operation of the servo loop (Figure 3) starts with the introduction of a dc signal through input circuitry designed for a wide range of voltages and currents. The scaled input is compared, within a high-impedance amplifier, to the feedback-potentiometer position voltage. The output of the am-

*US Patent 2,581,133 owned by General Radio Company.

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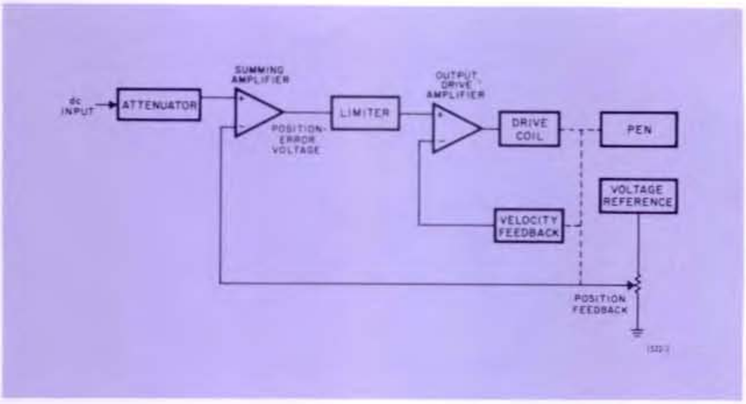


Figure 3. Block diagram of servo system.

plifier is the position-error voltage and is zero for the null position. If the input dc signal increases, a position-error voltage will be generated, which is fed through a limiter, to be compared with the velocity-feedback voltage. This difference is amplified to become the drive voltage for the pen motor assembly. The pen and wiper arm will move until the position-feedback voltage equals the input voltage, and the position-error voltage again is zero.

The velocity-feedback-voltage signal permits high feedback-loop gain at the null point (no velocity feedback), resulting in high static accuracy; dead-band error is less than 0.15%. The loop gain is effectively lowered during dynamic operations, keeping the position-feedback loop stable at high speeds.

Velocity feedback is derived from an unusual bridge technique (see Figure 4). Note that the output-drive amplifier serves both the linear servo motor and another fixed coil wound on a similar iron core. These two coils, together with two resistors, form a bridge network that can be balanced for ac as well as dc signals. When drive current is fed into this bridge network and the pen is stationary, the bridge is balanced; the output from the bridge and differential amplifier is zero. If the servo motor moves, the servo-motor coil develops a voltage proportional to the velocity; the fixed coil does not. Thus, there is a voltage output from the differential amplifier proportional to velocity.

Increasing the velocity feedback can simulate the action of a low-pass filter. The open-loop frequency response of this type of servo system is shown in Figure 5. As the velocity feedback is increased beyond the amount required for stabilization, the loop gain is decreased. The frequency at which the open-loop and closed-loop responses pass through 0 dB also is decreased. Thus the servo bandwidth is compressed by an amount proportional to any extra increase in velocity feedback. This action filters the effects of noise and undesired fluctuations in recording.

The position-feedback potentiometer is situated directly above the motor-drive coil. It is made of conductive plastic which assures low maintenance and extremely long life, greater than 20 million cycles, with negligible increase in noise or wear.

Full control of the chart drive, without mechanical clutches, is assured by use of three motors. Two induction motors, operating in a stalled condition, provide a constant holdback for chart supply and take-up and are instantly responsive to forward-or-reverse-direction instructions. The third motor, the stepper type referred to earlier, drives the sprockets that engage perforations in the chart, imparting a dependable motion to the chart at all times. Without the use of gears in the drive train, the recorder has eighteen chart speeds, from 0.5 s/in. to 20 h/in.

Operation of the stepper motor is controlled by a multivibrator, phase-locked to the ac power frequency, either 50 or 60 Hz. Its regulation and stability are similar to those of a synchronous motor. The maximum rate at which the motor will respond, without losing a step for instant start-stop operation, is 300 pulses per second. This rate corresponds to 2 inches/second (0.5 s/in.) chart speed. The actual rate of

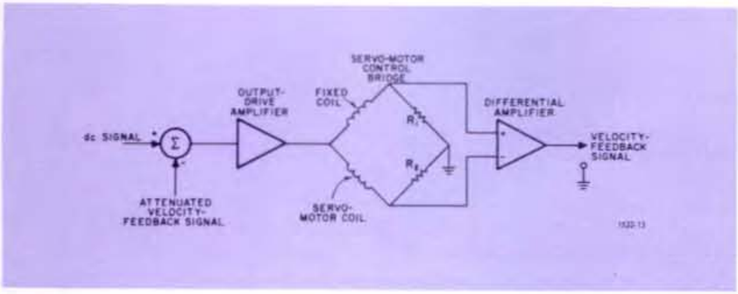


Figure 4. Schematic of velocity-feedback derivation.

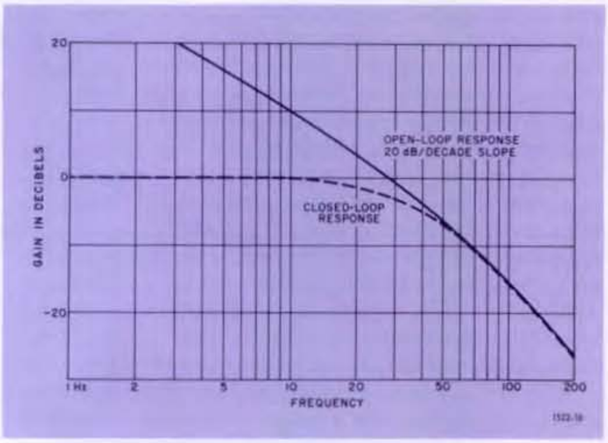


Figure 5. Servo-loop response.

rotation is locally or remotely controlled through integrated-circuit divider chains.

The output from the divider goes through a gating circuit to the stepper-motor driver circuit. This gate is controlled by a flip-flop for instant start-stop operation. The input to the flip-flop is capacitively coupled to both the front-panel switch and to the external programming input. Either input can be overridden by the other; the last signal determines the logic state. Logic memory is important for chart control; only a momentary pulse is necessary to set the state. Thus, when the photocell is used to change chart direction, the memory keeps the direction fixed after the mark is no longer under the photocell.

A forward-reverse flip-flop controls the direction of the stepper motor. Motor direction can be controlled locally or remotely

Output-control information, for use in system or automatic applications, is available from the recorder. As they pass over a photocell, the black timing marks on the chart (Figure 1) trigger output controls. Two photocells control as many as three output drives. When two in-line black marks pass the photocells at the same time, the paper motion is stopped. This method is used to signal the end of the chart roll; the ending of the roll is also verified by a printed note on the paper.

A switch is provided that lifts the pen and moves the chart at the fastest paper speed for scan operations; pen lift also occurs during scanning of the chart. In the AUTO position of the pen lift, the pen is dropped to the paper only in the RECORD mode and while the chart is moving.

Maximum and minimum limit switches are adjustable to any point along the chart. They provide electronic control for external operations such as sorting, inspecting, or sounding an alarm.

Situated next to the feedback potentiometer is a linear take-off potentiometer that can provide an external voltage proportional to the position of the pen. This voltage, and commercially available multiple integrated-circuit comparators, can establish limit-set positions controllable by an external potentiometer or voltage. The limit stops can be programmed externally, if a control voltage is used.

Another programming feature is pen blanking; when the servo system is turned off, pen motion freezes. The pen can be stopped anywhere on the chart during a scan operation, or during any operation that normally would force the pen off-scale during switching.

System Compatability

There are many possibilities for assimilating the GR 1522 into measurement and control systems. You are already aware of the functions performed by the GR 1921 Real-Time Analyzer (page 3) and of the value attached to permanent records of noise and vibration signals. The GR 1522, designed as a systems recorder, is a perfect complement to the GR 1921. A detailed description of the close working relationship between the recorder and analyzer may trigger readers' thoughts of other useful applications.

The 1926 detector stores the level of all the signal-input channels. It then feeds this information to the recorder at a rate that is controlled by the recorder's ability to accept and digest the data. Chart movement is continuous until the photocell control stops the recorder and changes channels in the detector. A fixed time delay of 0.1 second in the detector allows the pen to reach the signal level of the new channel before the detector sends a control signal to the recorder to start the chart. The pen then records the level of this channel at the fastest paper speed. The resultant plot has square corners, closely resembling the actual scope display. The start-stop-mark sequence repeats until the last channel selected by the analyzer is recorded. The pen automatically lifts, and the next chart moves rapidly into position. At this time, chart motion stops and the recorder does not function until commanded by the analyzer. It takes approximately 8 seconds to complete one chart and to advance to the next chart.

Table 1 shows the time required to make such a recording on three types of recorders: GR 1522, automatic X-Y recorder, and typical X-Y recorder.

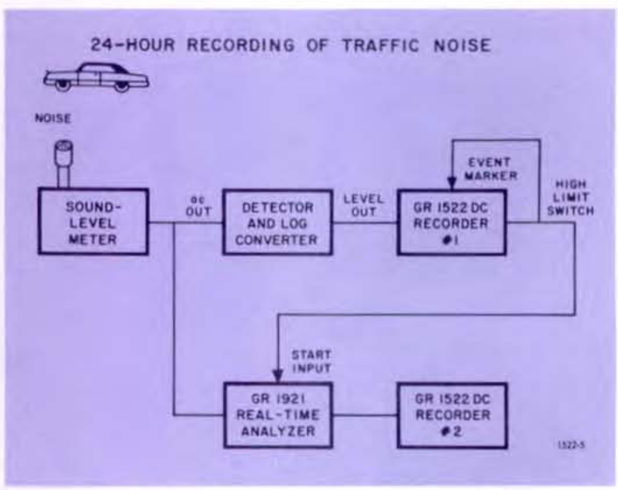


Figure 6. Typical long-time recording/analysis assembly.

Table 1

Recorder Type	Recording Time (seconds)	New Chart Advance Time (seconds)	Total Time (seconds)
GR 1522	6	1-2	7-8
Automatic X-Y	15	1	16
Typical X-Y	15	45	60

It is readily apparent that the 1522 takes one-half the time for each recorded spectrum, as compared with the automatic X-Y recorder, and is far superior to the typical X-Y unit.

Typical Applications

A practical application of the 1921/1522 combination, using two recorders (Figure 6), is the real-time spectrum analysis of band levels above a preset acoustical level. This might be a 24-hour study of traffic noise, aircraft fly-over noise, or similar patterns. During this interval we record, on Recorder No. 1, the average level of noise picked up and processed through a detector and log-converter circuit. Obviously, we don't want a spectrum every eight seconds because we would have 10,000 charts of recorded data in a 24-hour period. To save chart space and to compress accumulated data, the recorder is set at a slow chart speed. The 1921 analyzer is triggered by a GR 1522 limit switch to make a spectrum analysis of signals above a preset threshold level and also to identify the occurrence by printing an event mark on the Recorder No. 1 chart. After the GR 1921 is triggered, Recorder No. 2 receives instructions from the 1921 system to make the spectrum recording that corresponds to the indicated event. Upon completion of the study, a complete record is available, from Recorder No. 1, of the average sound level, and accumulated spectrum analyses of noise above a preset limit are presented by Recorder No. 2. All this can be accomplished without personnel in attendance.

Another system application involves the temperature-coefficient (Figure 7a) of a practically unlimited number of components. A scanner and the GR 1654 Impedance Comparator can measure 100 components consecutively, feed the information to the GR 1522, and plot the percent deviation from a standard at room temperature. A second plot can be

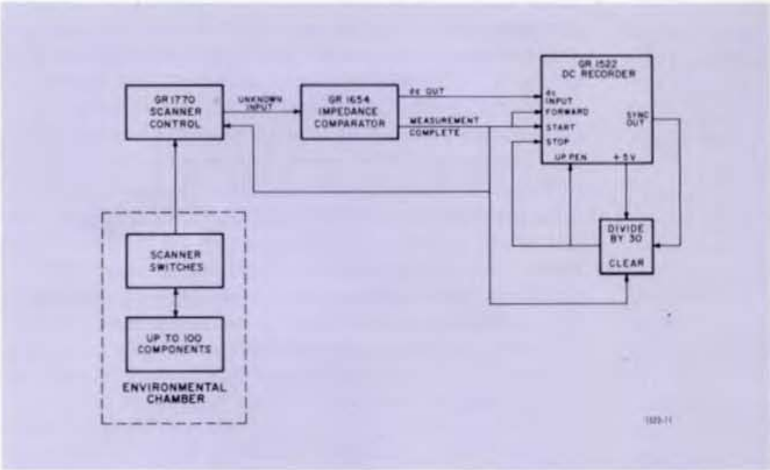


Figure 7a. Block diagram for multi-component temperature-coefficient measurements.

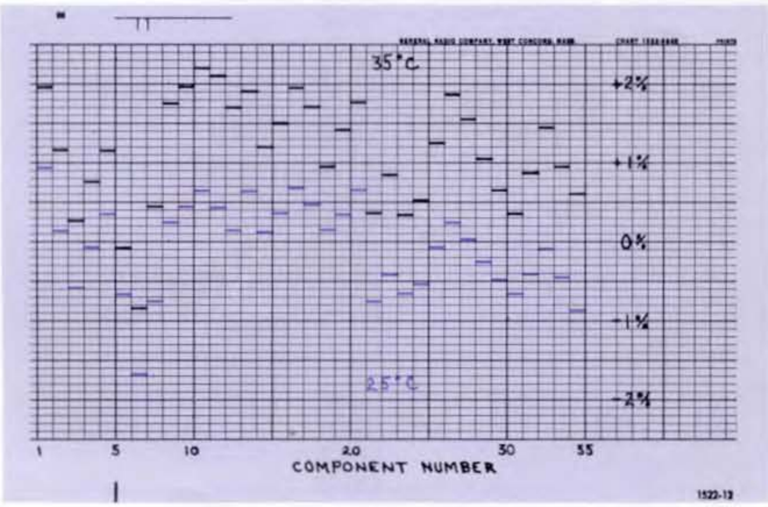


Figure 7b. Component test results. Chart motion is controlled by external divider circuit.

made using a second-color pen for a different temperature (Figure 7b). A measuring sequence is started by a switch on the scanner control. Upon switching to a component, the comparator waits for balance and then gives a measurement-complete command. This time delay allows the pen to move to the next component-recording position. The measurement-complete signal from the GR 1654 starts the recorder and paper motion. This particular chart paper has no control marks; the chart motion is controlled by external means. An external flip-flop divider divides, by 30, a sync signal used to drive the stepper motor. After 30 pulses, the chart has been moved a recording interval of 0.2 inch. At this instant, motion stops and the pen lifts. The same measurement-complete command is fed back into the scanner control, and the entire record cycle is repeated after the short delay period to achieve balance.

Remote programming for process control facilitates unattended recorder operation. A signal level, beyond preset limits, from a control transducer, triggers accelerated chart motion for greater detail, operates the scanner control to permit observation of other significant transducers, and then stops the process.

In another typical application, we can monitor a zener-diode voltage output. Here, both long-term drift and short-term noise are important. Photocell-control marks every 11 inches are used to trigger a flip-flop circuit that permits a long-term performance recording for 110 minutes (10 min/in.). This recording is followed by an accelerated recording of noise for 55 seconds (5 s/in.). Each recording is alternately and automatically presented in a continuous sequence of 11-inch charts.

To encourage specific application of the recorder to the GR 1921 analyzer, four different charts to cover various channel combinations and bandwidths used in the GR 1921 will soon be available from stock. Recorder channel widths of 0.208, 0.250, and 0.500 inch are available, dependent upon spacing of the photocell control marks on the chart edge. An unlimited variety of charts using various placement of the marks can be supplied on special order to customers.

M. W. Basch received his degrees at MIT (BSEE and MSEE-1958) and joined GR in 1958 as a development engineer in the Industrial Instruments Group and later transferred to the Acoustics/Signal Analysis Group. He has specialized in design of recorders and associated instrumentation. His memberships include IEEE, Tau Beta Pi, and Eta Kappa Nu.

ACKNOWLEDGMENT

The author gratefully acknowledges the help of M. C. Holtje in several critical areas and the support of R. P. Anderson for the power-supply design, of G. E. Neagle for the mechanical design, and of P. A. d'Entremont for the industrial design.

Complete specifications for the GR 1522 are included with this issue as a tear sheet, removable for insertion in GR Catalog T.

The Honorable Society

Long-time readers of the *Experimenter* will note with interest honors accorded two of their friends.

On May 1, 1969 Paul K. McElroy was awarded the annual Contribution Award of the IEEE Parts, Materials, and Packaging Group, at ceremonies conducted during the Electronics Components Conference of the Electronic Industries Association. "PK", now retired from GR, but hardly less active professionally than while he was with us, came to the attention of our readers for the first time in 1926. It was then he published a two-part article entitled "Design and Testing of Plate Supply Devices." He prepared a steady stream of written articles for almost forty years, terminating his output with "A 100-μF Decade Capacitor" in July, 1965. Our debonair retiree resides at 58 Gayles Road, Belmont, Mass.

The other recipient of honors was Dr. Arnold P. G. Peterson, GR consultant in the areas of sound and vibration. As part of ceremonies conducted to mark the twentieth anniversary of the foundling of the Audio Engineering Society, Dr. Peterson received the John H. Potts Memorial Award on October 23, 1968. The bronze medallion named for the deceased publisher and editor of "Audio Engineering" was given in recognition of outstanding achievement in the field of audio engineering. The citation noted Dr. Peterson's "many prominent contributions to the design of audio and acoustic instrumentation." Some readers may remember the first *Experimenter* article by APGP entitled "The Type 757-A UHF Oscillator" in the August 1941 issue; more may recall his latest article, "A Magnetic Tape Recorder For Acoustical Vibration And Other Audio-Frequency Measurements" in October 1966. It appears to be time for another article!

P. K. McElroy

A. P. G. Peterson

Type 1654 Impedance Comparator with its companion accessory, Type 1782 Analog Limit Comparator.

IMPEDANCE COMPARISON SPRINTS AHEAD

Mobility in measuring impedances, coupled with accuracy, wide range, and ability to control!

The Goal

Over the years, GR has tried to relieve the tedium and monotony of passive component inspection and testing by introducing functional-designed test equipment. For instance, we feel that technicians are able to remain alert for longer periods of time when test data are presented in simple GO/NO GO patterns. Eliminating the time normally required to read panel-meter indications also results in a saving in labor costs. In our estimation, the design of the new GR impedance comparator has successfully joined the concept of complex, programmed testing with the concept of simple, individual component measurements, at no sacrifice in accuracy or reliability.

The first comparison bridge engineered by General Radio for production testing was the manually-operated GR 1604 Comparison Bridge.[1] Several years later, it was followed by the GR 1605 Impedance Comparator,[2] incorporating greater precision and versatility, simultaneously indicating phase-angle and magnitude differences between two external test impedances over an extended frequency range, and requiring no bridge balancing. This bridge was semi-automatic in operation.

In 1964 the GR 1680-A Automatic Capacitance Bridge[3] brought with it a component test rate better than two per second. The inspection-rate bottleneck was broken, and the unit's acceptance by industry encouraged introduction of another digital unit - the GR 1681 Automatic Impedance Comparator System.[4] The customer received digital readout of phase-angle and magnitude differences, higher accuracy and resolution, and greater compatability with automatic component- and data-handling equipment.

Many analog-measurement operations and customers still existed. The need to extend and expand the design of analog comparators was the incentive to develop the latest type of GR comparator, which incorporates automatic features.

The Achievement

The GR 1654 Impedance Comparator supersedes the GR 1605 comparator and, with its companion accessory the GR 1782 Analog Limit Comparator, provides a semi- or fully-automatic system at about one-third the cost of a similar digital system. Measurement features include

- Operating range from 100 Hz to 100 kHz in four steps.

[1] Holtje, M. C., "A New Comparison Bridge for the Rapid Testing of Components," *General Radio Experimenter*, December 1952.

[2] Holtje, M. C., and Hall, H. P. "A High-Precision Impedance Comparator," *General Radio Experimenter*, April 1956.

[3] Fulks, R. G., "The Automatic Capacitance Bridge," *General Radio Experimenter*, April 1965.

[4] Leong, R. K., "The Automatic Impedance Comparator," *General Radio Experimenter*, June-July 1968.

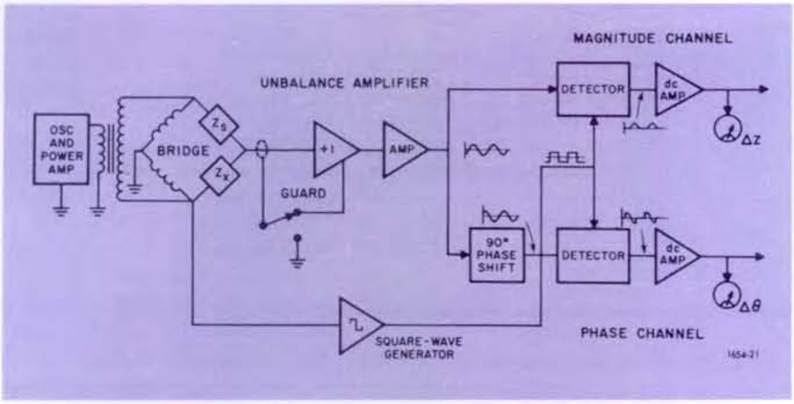


Figure 1. Block diagram of GR 1654 Impedance Comparator.

- Comparison precision to 30 parts per million.
- Impedance range from 2 Ω to 20 MΩ at 100 Hz.
- Capacitance range as low as 0.1 pF direct reading, with a modified substitution method.*
- Six deviation ranges from 0.1% to 30% full scale.
- Test voltages from 0.3 to 3 volts, for easy voltage-coefficient tests.
- Operating rate up to 4 units per second, in the automatic test mode.
- Positive indication of direction of overload.
- Indications on large panel meters.
- Analog output voltages.

*Refer to Circuit Notes at end of article.

The block diagram of the 1654 comparator, Figure 1, illustrates use of a tightly-coupled 1:1 ratio toroidal transformer as two arms of the bridge; the standard and test impedances complete the circuit. The unbalanced-output signal of the bridge is fed through a guarded circuit, which effectively reduces cable capacitance by three orders of magnitude. This permits measurement, with negligible error, of test items as remote as thirty feet. From the amplifier system the signal is passed to the magnitude and phase channels.

Input to the magnitude-channel phase detector is direct. Input to the phase channel first undergoes a 90° phase shift, before connection to the phase detector, in order to bring the test frequency signal in phase with the error voltage component due to any phase difference. The phase detector is fundamentally a switch operated in synchronism with the test frequency. Exact switching is controlled by a square wave derived from the zero phase signal of the bridge.

The rectified voltage is that component of the error voltage which is in phase with the controlling square-wave voltage. The detected output is fed to a stable dc operational amplifier which provides the required analog output voltage that is interpreted as magnitude difference or phase-angle differ-

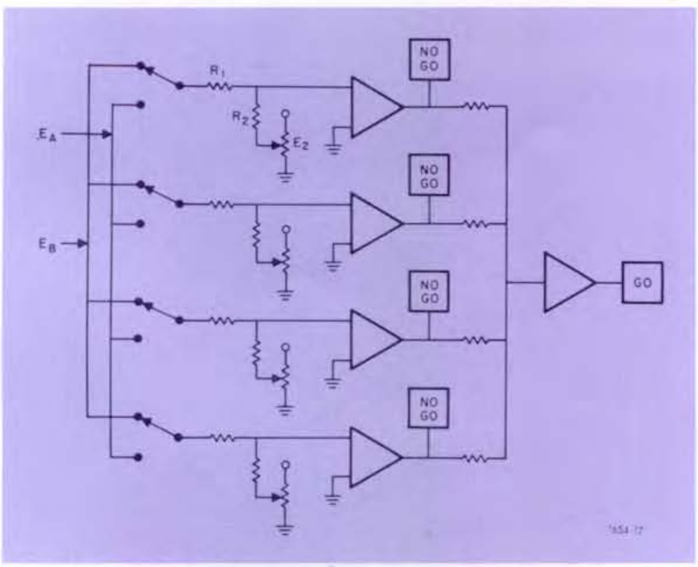


Figure 2. Block diagram of GR 1782 Analog Limit Comparator.

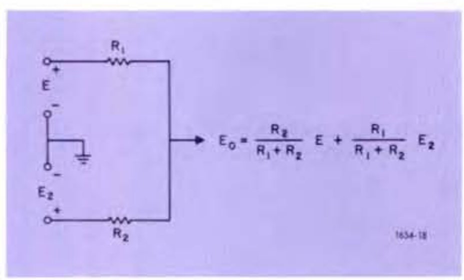


Figure 3. Schematic of limit comparison function.

ence between the test item and the standard.

The 1782 accessory unit (Figure 2) incorporates front-panel lamps to indicate GO or NO GO conditions. Each NO GO lamp indicates a single test limit, established as a preset voltage E_2. The input voltage E, derived from the magnitude or phase channels of the 1654, is compared with E_2, as shown in the schematic diagram, Figure 3. Unbalance voltage E_o, if it exists, is amplified by an operational amplifier with sufficient positive feedback to cause the amplifier to switch off or on. Its output triggers the NO GO lamp drivers; if the comparison is out of established tolerance, the NO GO lamps will light. Within-tolerance conditions for all limits will trigger the GO lamp. Optional relay-equipped models operate external automatic sorting devices.

Applications

On its own, the GR 1654 is a work horse on the inspection line or in the laboratory. Routine operations such as sorting, selecting, and adjusting passive components (R, L, and C), and any complex assembly of these components, are accomplished as quickly as you can make connections to the bridge. The GR 1680-P1 Test Fixture

R. K. Leong holds degrees from Northeastern University (BSEE-1960 and MSEE-1962). After two years' service in the U. S. Army as 1st Lt., coordinating instrumentation for upper atmosphere research, he joined GR in 1964 as a development engineer in the Low-Frequency Impedance Group. He has specialized in bridge and comparator design. He is a member of Tau Beta Pi and Eta Kappa Nu.

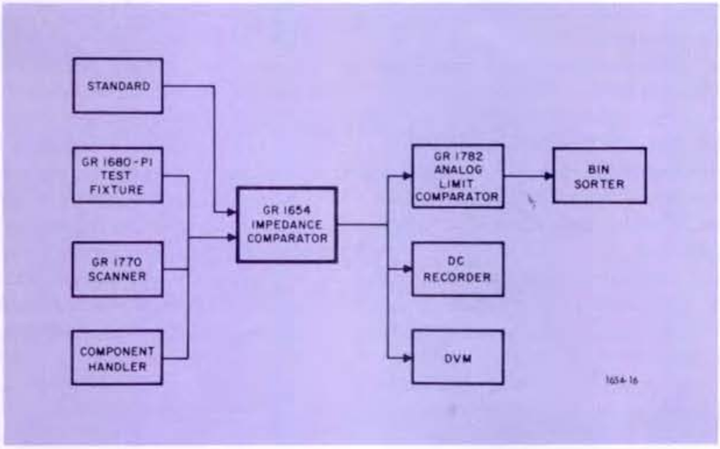


Figure 4a. Typical instruments and devices for automatic component-measurement system.

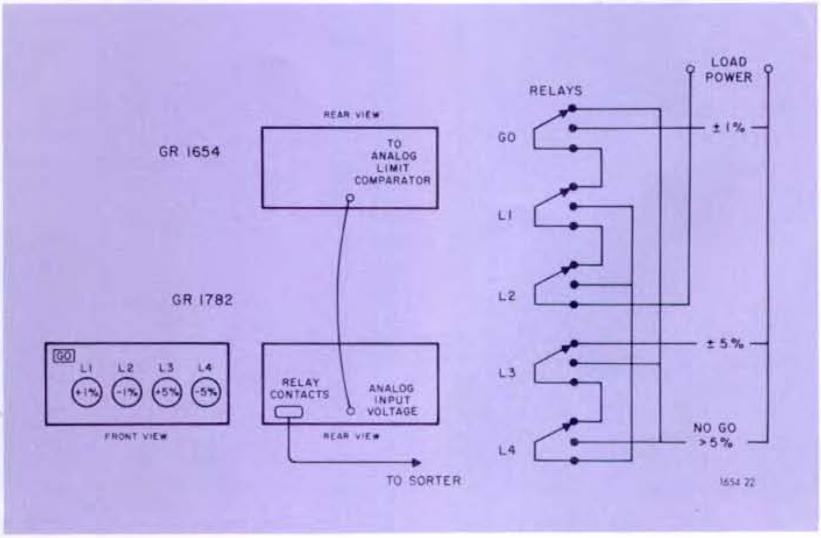


Figure 4b. Typical sorting system - schematic and relay interconnections.

is an ideal device for holding manually-inserted components. For faster sorting, the comparator and supporting instruments and accessories are an efficient automatic component-measuring system, as shown in Figure 4a. A typical sorting system is shown in Figure 4b.

The GR 1654 also is suitable for measurements of capacitors of extremely small values, low-loss capacitors, and dielectric samples; temperature-coefficient measurements of components tested within environmental chambers are a routine matter. Production testing of back-biased diodes and transistor-collector junctions is also routine; testing and adjusting of ganged capacitors and potentiometers for desired tracking within tolerances is another useful application.

—R. K. Leong

Complete specifications for the GR 1654 and 1782 instruments are available in the 1969 supplement to GR Catalog T.

Catalog Number	Description	Price in USA
	1654 Impedance Comparator	
1654-9700	Bench Model	$1300.00
1654-9701	Rack Model	1250.00
	1782 Analog Limit Comparator	
	Bench Model	
1782-9700	without relays	550.00
1782-9702	with relays	625.00
	Rack Model	
1782-9701	without relays	570.00
1782-9703	with relays	645.00

All prices subject to quantity discount.

CIRCUIT NOTES

The GR 1654 bridge circuit (Figure 5) measures impedance difference as a percentage of the average of the standard (Z_s) and unknown (Z_x) impedances. The real part of small phase-angle differences can be derived from the equation

$$\text{Re}\ \frac{E_o}{E} = \frac{|Z_x| - |Z_s|}{|Z_x| + |Z_s|} \quad (\text{When } \theta_x - \theta_s \leqslant 0.1 \text{ radian})$$

Introduction of nonlinear networks into the magnitude channel results in a linear indication of the impedance difference as a percent of the standard

$$\text{Re}\ \frac{E_o}{E} = \frac{|Z_x| - |Z_s|}{|Z_s|} \times 100\%$$

and reduces the number of scales for total measurement range. Measurements of R, L, and C furnish magnitude differences as percentages:

$$\frac{R_x - R_s}{R_s} \times 100\%, \quad \frac{L_x - L_s}{L_s} \times 100\%, \quad \frac{C_x - C_s}{C_s} \times 100\%$$

The imaginary part of small phase-angle differences can be obtained from the equation

$$\text{Im}\ \frac{E_O}{E} = \frac{\theta_x - \theta_s}{2}$$

For measurements of R, L, and C, the phase-angle difference indication is a measure of

ΔD of C and L, or ΔQ of R.

Measurement of low-value capacitances usually is limited by input-terminal capacitance C_{IN}. The capacitance effect is mitigated by the connection of the standard C_A, as shown in Figure 6. If the value of C_A is chosen as approximately $10 \times C_X$, or greater, the magnitude difference readout is proportional to $\dfrac{2C_X}{C_A + C_{IN}}$. For example, if C_{IN} is typically 1 pF and $C_A + C_{IN} = 200$ pF, for a meter reading of 0.1% the value of $C_X = 0.1$ pF. Note that the bridge is *direct* reading.

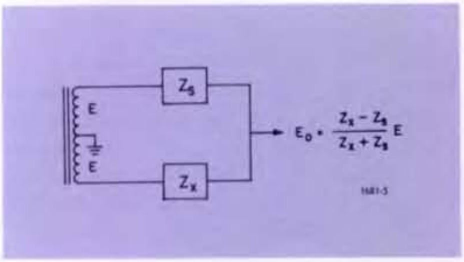


Figure 5. GR 1654 Bridge circuit.

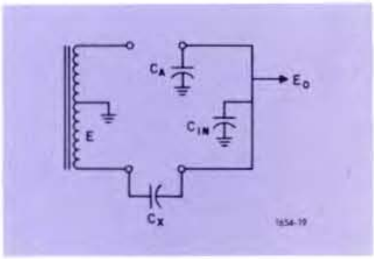


Figure 6. Bridge schematic for small-capacitance measurement.

Information Retrieval

Indexes for Volumes 41 and 42 (1967, 1968) of the *Experimenter* are available upon request to the editor.

VHF and UHF Attenuation Measurement to 140 dB[1]

This paper describes a system assembled for rapid measurement comparisons of attenuators with values ranging to 140 dB, with basic resolution of 0.01 dB and with maximum errors of 0.08 dB. A simplified system for measurements to 110 dB at certain selected frequencies is described also, as well as refinements permitting spot frequency measurements to 170 dB.

by S. Brown Pulliam

Introduction

The present-day design of commercially available standard-signal generators is highly sophisticated.* Quality assurance testing by the manufacturers must rely upon state-of-the-art techniques. An example is the testing of output-signal attenuators, which control the signal level over a range of 0 to 140 dB in 10-dB steps. Maximum allowable error is 0.1 dB per step; maximum cumulative error is 0.5 dB. The frequency range for these requirements is from dc to 500 MHz. This article deals with the test measurement system that was designed to assure instrument performance to specifications.

The System Choice

R. W. Beatty of NBS has classified the various attenuation-measurement systems;[2] our system falls in the general category "Direct Substitution." Several techniques are available under this system, including series and parallel substitution. The major advantage of the series substitution technique is that, as the test attenuator is adjusted in level, the reference attenuator is similarly adjusted but in the opposite direction. The total circuit attenuation therefore remains constant and, the signal level into the detector being relatively constant, nonlinearity is not a problem. In the parallel-substitution technique, the test signal is split into two paths, one through the test attenuator and the other through the reference attenuator, and is recombined in a detector. This has the advantage that any instability in the output level of the source, or in the gain of the detector, has no effect upon measurement accuracy. However, in order to make the system work, either an auxiliary rf switch or a precision phase shifter is required.

Most commercially available attenuation-measurement systems use an i-f substitution method in which the reference attenuator is calibrated at one frequency only. These systems are limited to a maximum single-step range of approximately 100 dB due to the dynamic range of the mixer.

There is no definite basis for a choice between the series- and parallel-substitution techniques; however, with sufficient level stability of signal source and detector, the series-substitution technique is simpler to implement.

Any practical system for measuring large values of attenuation involves detection of a weak signal. Most engineers are aware of the fact that increased sensitivity involves narrowing of the receiver bandpass but may overlook the fact that some receiver systems contain elements which have a "threshold effect." One such element – the second detector of the usual single-conversion superheterodyne receiver – will lose an applied signal if it is buried in random noise,[3] and no amount of post-detection filtering can retrieve it. Fortunately, in most attenuation-measurement applications the available test signal is large enough, using conventional superheterodyne techniques, to obtain detection resolution equal to that obtainable by very sophisticated synchronous detection schemes.

System Design Considerations

Little of the literature concerning attenuation measurement systems is devoted to the problem of ease or rapidity of readout versus system sensitivity. Measurement range can be increased by narrowing the bandwidth, and, similarly, resolution can be increased by lengthening the measurement time. Decisions must be reached on what resolution or precision is required in a particular measurement and what maximum attenuation must be inserted in the measurement channel. The remaining significant variables are time, generator power, and detector noise figure. Time

*Examples are General Radio Types 1003-A and 1026-A.

[1] Abstracted from paper of same title scheduled for publication in the March, 1969 issue of the *IEEE Transactions on Instrumentation and Measurement.*

[2] Beatty, R. W., "Microwave Attenuation Measurements and Standards," *NBS Monograph 97,* April 1967.

[3] Downing, J. J., *Modulation Systems and Noise,* Chapter 4, Prentiss-Hall Inc., 1964.

is the most difficult variable to resolve and involves factors which are quite subjective. An experimental conclusion reached was that peak-to-peak meter jumps should be less than 0.02 dB in a measurement interval less than 5 seconds. This in turn required an equivalent signal-to-noise ratio in the receiver of at least 50 dB in a 1-Hz bandwidth. Assuming a total attenuation of 140 dB to be measured, the ratio of input signal to output noise is 190 dB. Signal generator power was calculated to be +21 dBm, considering a theoretical available thermal noise level of −174 dBm for a 1-Hz bandwidth (300 K operating temperature) and assuming a receiver noise figure of 5 dB. Available powers from the signal generators under test were +27 dBm (Type 1026-A) and +22.5 dBm (Type 1003-A), presenting no problem.

The signal versus noise requirements outlined above, combined with the need to avoid "threshold-effects" losses, resulted in selection of a pre-detection bandpass of 1 kHz. This immediately imposed stringent frequency-stability requirements upon the measurement system.

Working in the uhf spectrum with an LC tuned signal generator, which could easily drift more than 1 kHz, required an auxiliary phase-lock loop. The dual phase-lock system shown in Figure 1 was developed for the uhf tests. A synchronizer phase locks the source frequency, and a second loop locks the second i-f signal to a 50-kHz crystal-controlled reference frequency. This permits optional use of a wave analyzer or selective voltmeter if extension of the attenuation measuring range to 170 dB is required. A lock-loop bandwidth of 500 Hz is more than adequate to correct for any frequency drift in the crystal-controlled converter, when the 50-kHz phase-detector output is used to control the synchronizer reference frequency.

The basic readout instrument of the attenuation-measurement system is a 1-dB expanded-scale meter at the i-f amplifier output. It was slightly modified to provide a one-second time constant. An auxiliary readout recorder has become an integral part of the measurement system because it reduces operator fatigue considerably.

S. B. Pulliam joined the engineering staff of General Radio in 1959 as a member of the Signal-Generator Group. He had received the BS degree in Physics from Union College in 1951, served in the US Navy for three years as Electronics and Communications Officer, and also had been associated with the Naval Ordnance Laboratory as Electronic Scientist. Mr. Pulliam is a member of IEEE.

The vhf-measuring system developed is identical to the original uhf system with the exception that the converter is replaced by a 79-MHz converter. No synchronizer is required in this system. The frequency stability of either signal generator is more than adequate to stay within the 500-Hz locking range of the secondary loop, and the dc control is connected directly to the signal-generator varactor terminals.

System Performance Evaluation

Tests of the drift in level of the combined signal generator and 30-MHz amplifier indicate a stability of ±0.01 dB, for a time interval of 20 seconds. Further tests have indicated drift does not exceed ±0.015 dB for intervals up to 10 minutes.

Signal levels at the receiver input as low as −143 dBm produce successful phase locking. For useful measurements at this level, however, a second i-f bandpass on the order of 10 Hz is required. This may be provided by the wave analyzer as shown in Figure 1.

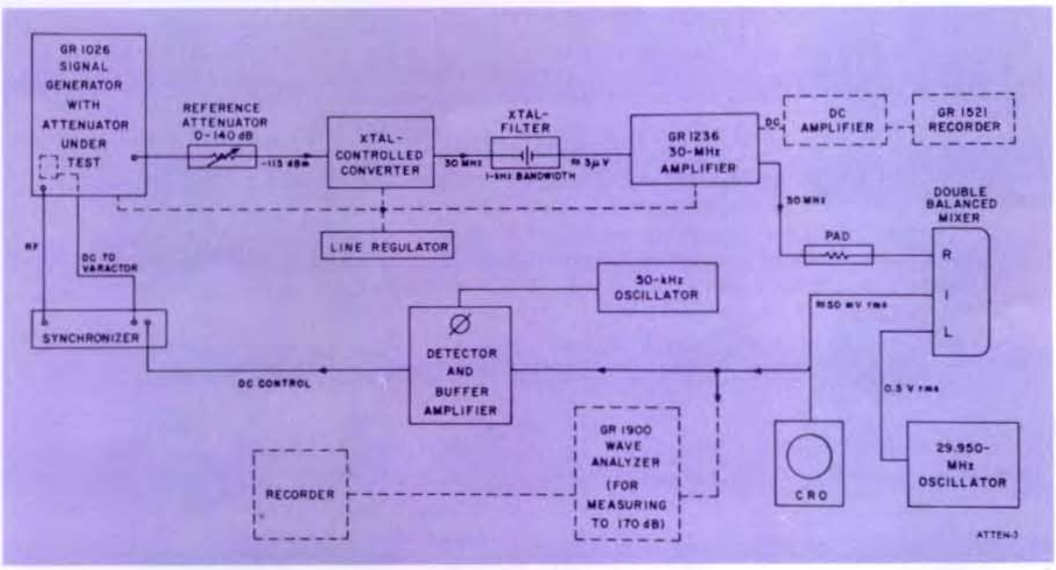


Figure 1. Block diagram of attenuation-measuring system.

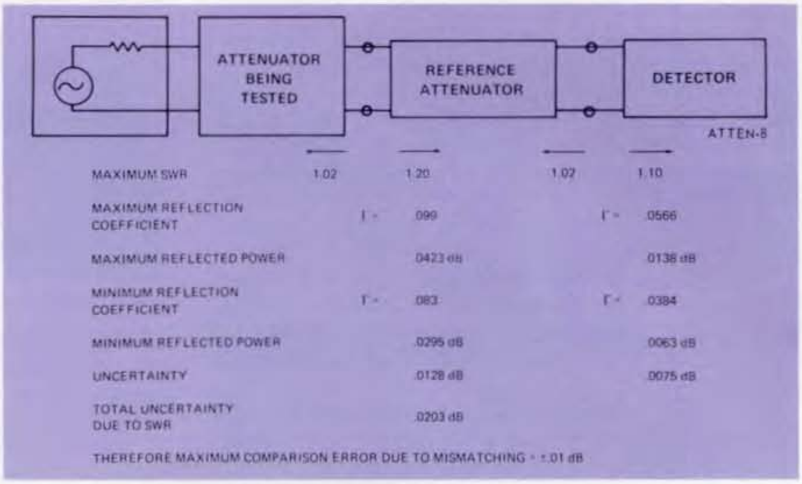


Figure 2. Significant reflection errors for interior steps of test and reference attenuators.

Figure 2 presents the interface reflections for the interior 100 dB of the attenuators and results of calculations for the interior steps. Attenuation comparison error for all interior steps is calculated to be ±0.01 dB. For the extreme steps an additional reflection pair between the generator and the attenuator being tested causes a maximum error that is calculated to be 0.06 dB.

Table 1
Summary of comparison errors

Source	Interior Attenuator Step	Extreme Values to 120 dB	to 140 dB	to 160 dB
Recorder resolution and system drift	±.01 dB	±.01 dB	±.01 dB	±.01 dB
SWR	±.01	±.06	±.06	±.06
Noise fluctuation	0	0	±.01	±0.2
Meter error (1000)	–	–	–	±0.3
Worst-case error	±.02 dB	±.07 dB	±.08 dB	±0.57 dB

Much attention was directed to rf leakage around the attenuator path because of the large amounts of attenuation involved. Exceptional care was taken to shield the source from the detector. A leakage signal which is coherent with the detected signal can cause an error of 0.01 dB if its level is 60 dB below that of the detected signal. A measurement at the 140-dB level would require shielding effectiveness of 200 dB. In addition to the requirement for well shielded signal sources and detectors, interconnections must also be well-shielded by solid-sheath coaxial cable. Tests made with the system described in this article showed that in no case was leakage found to be a disturbing factor for measurements to 140 dB.

The summary of the errors, shown in Table 1, indicates a worst-case comparison error of 0.08 dB for attenuations to 140 dB, and 0.57 dB to 160 dB.

Editor's Note – Copies of Mr. Pulliam's paper will be available from General Radio Company, 300 Baker Avenue, West Concord, Massachusetts 01781.

Seminar Scheduling

A new seminar has been added to General Radio's Customer Seminar Program. Its title is Real-Time Sound and Vibration Measurements and it covers theory and application of the GR 1921 Real-Time Analyzer, input/output equipment, and analyzer/computer systems. Two 3-day seminars were scheduled during June at the Concord facility. This fall, the seminars will be repeated both at Concord and in several major cities. Contact your local GR District Office for more information.

Recent Technical Articles by GR Personnel

"Instruments and Techniques of Sound Measurements," W. M. Ihde, *National Safety News*, December 1968.*

"Simplified Pure-Tone Audiometer Calibration," R. P. LaJeunesse, *National Hearing Aid Journal*, March 1969.**

"Tone-Burst Testing of Underwater Transducers," D. M. Lloyd, *Undersea Technology*, April 1969.*

"Recent Advances in Coaxial Components for Sweep-Frequency Instrumentation," T. E. MacKenzie, *Microwave Journal*, June 1969.

"Ratio Transformer Techniques for Precise Gain and Phase Measurements," T. J. Coughlin, *Electronic Instrument Digest*, June 1969.

*Reprints available from General Radio.
**Expanded Instrument Note (IN114) available from General Radio.

Type 1160-RDI-1B Digit Insertion Unit.

FASTER SWITCHING FOR 1160-SERIES SYNTHESIZERS

Why and How

The General Radio 1160-series of frequency synthesizers is significantly improved, primarily in programmed applications, with the introduction of the GR 1160 RDI-1B Digit Insertion Unit. It offers higher switching speed and practically unlimited switch life by replacing reed relays with semiconductor diodes. It retains and expands the remote-control capabilities of the original 1160 RDI-1 unit.[1]

The use of synthesizers has increased considerably since their introduction to the fields of electronic measurement and instrumentation. The synthesizer transfers the accuracy and stability of a frequency standard to a broad range of frequencies with almost unlimited resolution.[2] Additional capabilities, such as electronic frequency selection from an external program, have opened up many applications not anticipated previously and have caused a demand for increased switching performance. The new digit insertion unit is expected to meet this demand.

The modular arrangement of the General Radio synthesizer family allows us to use identical digit insertion units in a chain to control the successive digits in the number defining the output frequency.[3] One group produces frequency increments from 100 kHz to 0.01 Hz, depending upon its position in the chain. This group of units is identified as RDI-1 (the letter R indicates remote-control capability). The early programmable units contain reed relays. Switching is relatively slow (2 ms) and typical life approximates 10^8 operations. It should be noted that a number of this magnitude could be exceeded in relatively short time in an application involving continuous high-speed switching.

The new RDI-1B unit eliminates both disadvantages. Switching time now is less than 200 μs (an order of magnitude improvement), and life expectancy is now almost unlimited.

Figure 1 shows a transition from 90 kHz to 20 kHz. No phase discontinuity is visible with an oscilloscope sweep rate of 50 μs/cm. It is important to note that this kind of presentation permits no assessment of the accuracy of the newly-established frequency immediately after switching. Synthesizer systems employing any kind of narrow-band devices after the point of frequency switching will have stabilizing times and frequency errors which may not be insignificant in relation to the normally high accuracy of synthesizers. Higher-resolution measurement techniques are therefore required.

Figure 2 shows an example of the method that was employed to obtain repetitive switching sequences from 900 kHz to zero frequency, using a dc-coupled phase detector for observation. From inspection of the oscillograph, it is evident that the proper phase relationship is obtained in most cases in less than 100 μs (horizontal scale is 200 μs/cm). The lower trace represents the programming signal (voltage on "0" programming line).

Precision sweep capabilities, with the continuously-adjustable decade, account for numerous applications of GR synthesizers in network characterization. In addition to these capabilities, it is now possible to simulate a

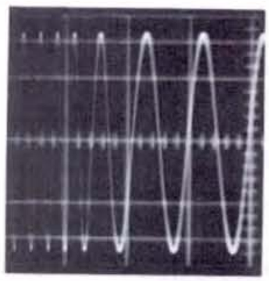


Figure 1. Switching transition, 90 to 20 kHz.

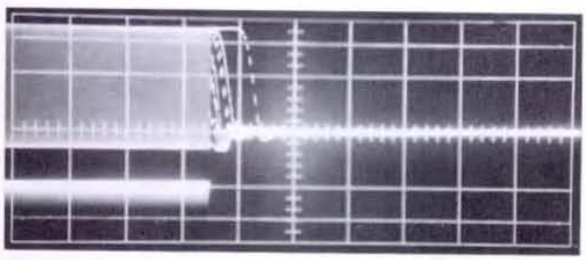


Figure 2. Switching sequences - 900 kHz to zero Hz.

[1] Lohrer, G. H., "Remote Programming for GR Synthesizers," *General Radio Experimenter*, May 1965.

[2] "Applications for Coherent Decade Frequency Synthesizers," General Radio Company, October 1968. Copies available on request.

[3] Noyes, Jr., A., "Coherent Decade Frequency Synthesizers," *General Radio Experimenter*, September 1964.

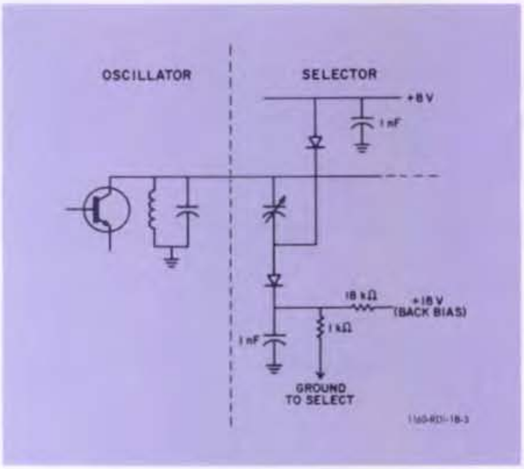


Figure 3. Simplified diagram of diode-switched oscillator - one section of selector only.

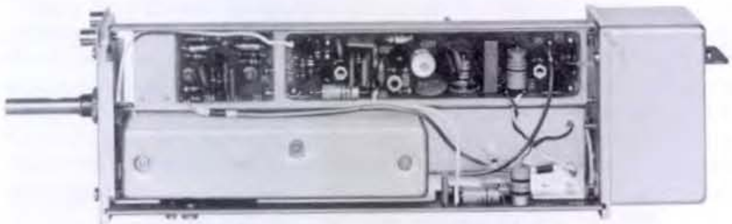


Figure 4. Interior details of RDI-1B unit.

sweep by remotely controlling frequency in suitably small increments, with no restrictions imposed by relay-contact life. The GR RDI-2, -3, and -4 insertion units already are programmable by the electronic control in steps of 1 MHz or 10 MHz.

The new RDI-1B unit replaces both the DI-1 and the RDI-1 units. Synthesizer models without remote frequency programming are still available; in the future they will be supplied with the RDI-1B module but without remote-control filtering. Filter plugs and remote cables are available separately, on special order, for remote-control programming, if desired at a later date. A 10-line code is normally provided but 1-2-4-8 BCD coding is available on special order.

Design Details

An important aim of the new design was circuit simplification. This was accomplished by using a simple transistor input mixer, a new phase-lock circuit design, and a divide-by-ten circuit differing considerably from the previous design. A higher switching speed was desired in the phase-lock loop that selects the proper digit from a picket fence of frequencies. We obtained it, and also an economic advantage, by employing silicon diodes to connect the different frequency-control capacitors and by redesigning the phase-lock loop for highest speed consistent with proper attenuation of unwanted frequencies.

Figure 3 shows a simplified schematic of the digit oscillator and selector. Two diodes are used per digit. They provide a double path for the rf current to ground in the "on" state and obviate the need for an rf choke in the "off" state. The energized end of the capacitor is isolated by means of the diode back bias. The digit oscillator is part of the active filter that selects the proper digit from ten available frequencies.

To illustrate further the changes in design, we divide 50-51 MHz by 10 with an injection-locked oscillator, directly synchronized by the high-frequency signal. A circuit design, which guards against free-running of the oscillator, causes the output amplifier to turn off automatically when an input signal has not reached the divider.

Improvements in reliability and in performance of the synthesizer-digit insertion unit combination have been attained through reduction of some spurious outputs and susceptibility to magnetic disturbances. Most significant, however, is the reduction of switching time from 2 ms to 200 μs.

You will note, from Figure 4, that the outer appearance of the RDI-1B is very similar to that of the RDI-1 and DI-1 units, indicating interchange compatability. The outward appearance, however, is all that is similar. Four circuit boards of identical dimensions greatly facilitate semi-automatic alignment and testing. Functional compartmentizing and shielding have improved serviceability and performance of the unit.

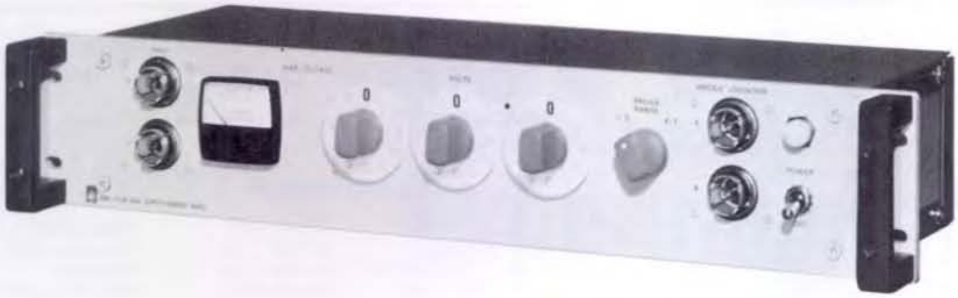
—G. H. Lohrer

G. H. Lohrer is a 1952 graduate of Technische Hochschule (Dipl Ing EE), Karlsruhe, W. Germany. His first employment was with G. Lorenz A. G.; he joined Canadian GE in 1953, Philips Electronics in 1955, National Company in 1959, working in various capacities in the VHF-UHF field. In 1961, he joined General Radio where he has specialized in synthesizer work. A senior member of IEEE, he is also a member of the Ontario Professional Engineers Association.

Complete specifications for the GR 1160 RDI-1B are available in the 1969 Supplement to General Radio Catalog T.

Catalog Number	Description	Price in USA
	1160-RDI-1B Digit Insertion Unit	
1160-9485	less filter plug	$455.00
1160-9480	with filter plug	505.00

All prices subject to quantity discount.



2995-9158 Bias Supply.

The measurement of capacitors under dc bias has become more than an occasional task for the Type 1680 Automatic Capacitance Bridge, and we have responded to this and other requirements with a companion bias supply, the 2995-9158.

This supply provides a bias voltage adjustable from 0 to 50 volts* in increments as small as 0.1 volt. Voltages can be remotely programmed or manually set, and a panel meter serves to indicate the output level. The 2995-9158 also protects the GR 1680 from large current surges by instantaneously limiting the current whenever a charged or shorted capacitor is inadvertently connected for measurement.

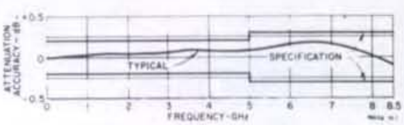
An additional feature of this bias supply is the very low series impedance it presents to the bridge test signal. This eliminates the need for corrections to the GR 1680 readings on all but the highest capacitance range. Engineering development was by R. P. Anderson, Assistant Group Leader, Impedance and Networks Group.

*Several supplies can be used in series to permit a combined bias voltage of 150 V.

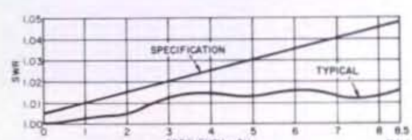
Voltage Range: 0 to 50.0 V.
Accuracy: (0.2% + 10 mV), typical.
Stability: (0.1% + 1 mV), typical for 8 hours.
Impedance: Less than 0.2 Ω from 0 to 1 kHz.
DC Current Limit: 40-mA positive, 10-mA negative.
Transient Current Limit: Less than 100 mA within 2 μs.
Added 3-Terminal Capacitance: Less than 1 pF.
Programming: 100 Ω/V.

Catalog Number	Description	Price in USA
2995-9158	**Bias Supply**	**$800.00**

All prices subject to quantity discount.



Type 900-G6, -G10 Attenuation Accuracy.



Type 900-G6, -G10 SWR.

GR 900-G6 and GR 900-G10 Precision Fixed Attenuators are equipped with GR900® connectors and have values of 6 dB and 10 dB respectively. These attenuators are much lower in SWR and have a more uniform attenuation over a wide frequency range than units previously obtainable for 50-ohm lines.

GR 900-D20 Adjustable Short Circuit is a coaxial sliding short circuit equipped with the GR900 connector, usable as a tuning and matching element or as a reactance standard.

GR 900-LK10 Precision Adjustable Line is a line stretcher with very low SWR and uniform characteristic impedance and insertion loss, equipped with GR900 connectors.

Complete specifications for the four units above are available in the 1969 Supplement to General Radio Catalog T. Engineering development was by John Zorzy, Group Leader, GR Engineering Department.

Catalog Number	Description	Price in USA
	Precision Fixed Attenuators	
0900-9850	900-G6 (6 dB)	**$185.00**
0900-9851	900-G10 (10 dB)	185.00
0900-9430	**900-D20 Adjustable Short Circuit**	160.00
0900-9570	**900-LK10 Precision Adjustable Line**	245.00

All prices are subject to quantity discount.

Type 900-G10

Type 900-D20

Type 900-LK10

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