NRL Report 4664

VIDEO-PRESENTATION ANALYZER
50 KC TO 10 MC

F. T. Griffin

Countermeasures Branch
Radio Division

March 8, 1956

NAVAL RESEARCH LABORATORY
Washington, D.C.

CONTENTS

ABSTRACT

Video-frequency spectra data, from the modulating source as well as the incorporated amplifier, of an electronic countermeasures equipment are often required in determining the equipments' effectiveness. Previous facilities utilized for video analysis were characterized by slow operational point-to-point sampling procedures, and by uncertainty of resolution figure due to possible Q variation of each of the multiple-tune circuits used in the analyzer.

After a brief discussion on the merits and limitations of tuned-circuit type analyzers as contrasted with superheterodyne types, a quasi-instantaneous video analyzer which was developed principally for the rapid analysis of modulation spectra is described. The instrument presents the frequency analysis of a complex voltage visually, with an alternative provision for instrumental analysis by manual control. It is essentially a double-conversion superheterodyne receiver without preselection. A magnetically controlled local oscillator is used as a sweep for visually displaying the complete spectrum from 50 kc to 10 Mc, on a cathode-ray tube. In addition, a manually tuned local oscillator is used in conjuction with a thermistor bridge indicator for a point-by-point analysis.

The superheterodyne principle permits a suitable, constant, and controlled resolution value to be used, and allows a visual analysis to be made in 1/30 second by sweeping the local-oscillator frequency over the desired frequency band at a 30-cps repetition rate.

A relation exists between the three factors — the video frequency band to be analyzed, the time required to give a complete analysis, and the final resolution bandwidth of the analyzer. For a video frequency band of 50 kc to 10 Mc, and an analysis time of 1/30 second, the minimum permissible resolution bandwidth is 17.5 kc. The resolution may be increased by using a smaller video band to be analyzed, or by increasing the time permitted for a complete analysis.

The resolution bandwidth of the instrument is approximately 30 kc for either visual or manual type of presentation. The input probe presents relatively light loading to the voltage source to be analyzed. A three-step 20-db attenuator in the probe provides for inputs up to 280 volts peak to peak. The sensitivity for visual analysis is 880 microvolts rms, based on the signal required to exceed the inherent noise by 3 db. The manual point-by-point sensitivity (by the same standard) is 38 microvolts rms.

With this instrument an analysis can be made more quickly than with equipment employed in this section heretofore.

The instrument has given satisfactory laboratory performance for a period of two years, and could serve as a basis for design of a service equipment.

PROBLEM STATUS

This is an interim report; work is continuing on this problem.

AUTHORIZATION

NRL Problem R06-09
Project Nos. NR 506-090 and NE 071-240-8
BuShips Project S-1249

Manuscript submitted November 1, 1955

VIDEO-PRESENTATION ANALYZER
50 KC TO 10 MC

INTRODUCTION

The present methods of studying the amplitude and frequency distribution of various video signals, using point-by-point analysis, are tedious and slow. One of these methods (Fig. 1a) is based on the evaluation by a suitable detector of the voltage associated with particular frequency components, selected by a resonant inductance-capacitance circuit of controlled Q or certain forms of bridged T networks, inserted across the signal source being investigated. Later versions of analyzers, generally in the audio-frequency range (Fig. 1b) have been made using the heterodyne principle. In these instruments the local oscillator is varied by hand, making it possible to change the response frequency while using a fixed-frequency filter of predetermined characteristics (the intermediate-frequency amplifier) as the resolving element.

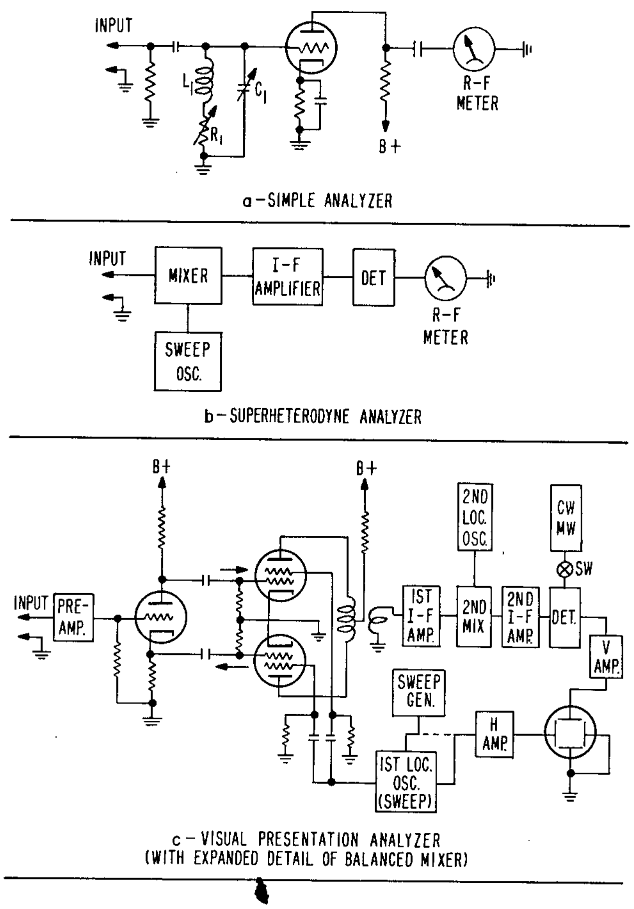


Fig. 1 - Simplified analyzers

These methods of analysis are slow in practice and a search for quasi-instantaneous sampling video analyzers, which would present the frequency on one axis of a cathode-ray tube and the amplitude of the complex wave components on the other axis, was begun in March 1952. No such analyzers, in the frequency range desired (50 kc to 10 Mc), had been described to the writer's knowledge at that time, and plans were made to design a preliminary equipment to permit study of the feasibility of such a proposal.*

GENERAL ANALYZER CONSIDERATIONS

The desirable characteristics of the superheterodyne analyzer, which has the ability to tune over a wide frequency range with constant resolving bandwidth and which can be designed to meet specifications closely in Q and bandwidth, seem to be preferable to the single-tuned type which, in spite of its desirable wide dynamic range, can be controlled only with difficulty over a limited frequency band. The superheterodyne may be of the double-conversion type as in Fig. 1c, to provide extremely narrow i-f bandwidth (high resolution) and reduce spurious response problems of certain types as well. A disadvantage may be claimed in that an analyzer with an i-f amplifier coupled to an inherently noisy balanced modulator, as required in this case, has less dynamic range than the single-tuned type of analyzer.

The frequency band to be analyzed, the amount of resolution desired, and the minimum rate at which the sampling may be made are interrelated to the circuit elements that may be selected for incorporation in an analyzer design. The selection of an intermediate-amplifier frequency is restricted by the highest value of the video band that it is desired to analyze, and should be so high that first-order harmonics of the video signal do not affect the i-f amplifier directly without benefit of conversion in the mixer. The second i-f amplifier frequency should be selected so that the desired resolution may be obtained, with due consideration of the characteristics of available commerical components such as transformers.

A nominal limit of 10 megacycles was chosen as the upper frequency limit for an analyzer that would be most useful for purposes encountered in electronic countermeasure studies. With this upper limit and a lower limit of 50 kc, and a sweep frequency high enough (above 20 cycles per second) to prevent visual flicker from the usual cathode-ray screen, an investigation of the resolution permissible under such specifications was begun. The desirability for high resolution leads early to a determination of the limits that cannot be exceeded. The assumption that the time a given frequency component (as determined by the modulation product of the signal and the local oscillator) remains within the bandpass of the final i-f amplifier will not be exceeded by the time constant of the final i-f amplifier, would apply here as with other types of frequency sweeping apparatus, e.g., sweeping oscillators for aligning receivers. This sweeping rate is determined by the type and frequency of the local oscillator sweep which, in this case, is linear with negligible flyback time.

In terms of total bandwidth to be surveyed, time of survey, and resolution bandwidth, the following symbols may be used:

Let:

B = total bandwidth to be surveyed (50 kc - 10 Mc)

T = total time of survey (1/30 sec)

*A description of an audio-frequency analyzer using cathode-ray presentation was given by S. V. Soanes in "Some Problems in Audio Frequency Spectrum Analysis," Electronic Engineering, 24:268, June 1952.

Δf = resolution bandwidth of final i-f amplifier

$n = \frac{B}{\Delta f}$ number of resolved bandwidths Δf in spectrum to be surveyed

$\Delta t = \frac{1}{\Delta f}$ time constant of final i-f amplifier

Equating the time constant of the resolving i-f amplifier to the time permitted to survey a single Δf:

$$\frac{T}{n} = \Delta t = \frac{1}{\Delta f} \tag{1}$$

Eliminating n,

$$T = \frac{B}{\Delta f^2} \tag{2}$$

The sweep time of 1/30 sec, selected as a submultiple of the main period (1/60 sec) has advantages in avoiding hum problems.

Then

$$\Delta f = \sqrt{\frac{B}{T}} = \sqrt{\frac{10 \times 10^6}{1/30}} = 17.48 \times 10^3 \text{ cycles} \tag{3}$$

Using 17.5 kilocycles as a minimum permissible resolution bandwidth, the local sweep oscillator may be designed to sweep the 10-Mc band in 1/30 sec. The problem as stated assumes the use of a linear sweep, but obviously other sweeps (e.g., logarithmic) could be used. It is necessary only to ascertain that the sweep rate, df/dt does not exceed the critical value Δf^2 at any time. It follows that once the resolution bandwidth Δf is determined, a practical superheterodyne analyzer circuit may be designed. This basic type of system seems best suited to fulfill the countermeasures requirement.

ANALYZER DESIGN, GENERAL

The final design of the video-presentation analyzer follows the principles of the superheterodyne with modifications. The block diagram (Fig. 2) shows both a sweep and a manual first oscillator. The sweep oscillator provides the quasi-instantaneous sampling of the video signal, and is associated with the visual presentation; the manually tuned oscillator provides a means of sampling a given portion of the video signal, and is associated with the bridge and meter presentation for absolute level determination. A balanced mixer is used to prevent the local-oscillator signal from entering the first i-f amplifier directly.

The operation of the analyzer will be outlined with reference to Fig. 2. The complex frequency to be studied (between 50 kc and 10 Mc) is applied through a phase divider to the grids of the balanced mixer. The local oscillator, sweeping at a 30-cycle rate, mixes with the complex frequency, giving at any instant the sum and difference frequencies. The difference frequency is accepted by and amplified in the first i-f amplifier at a center frequency of 50 Mc. This signal is then mixed with the second local oscillator. The difference frequency from this mixing is again amplified in the second i-f amplifier at a center frequency of 1 Mc and detected. The detected output is applied to the vertical plates of the oscilloscope for visual presentation. The sweep for the horizontal deflection of the oscilloscope is derived from the sweep which drives the sweeping local oscillator and is in synchronism with it. A relation is thus established between the video frequency under study, the local-oscillator frequency, and the position on the horizontal trace of the oscilloscope.

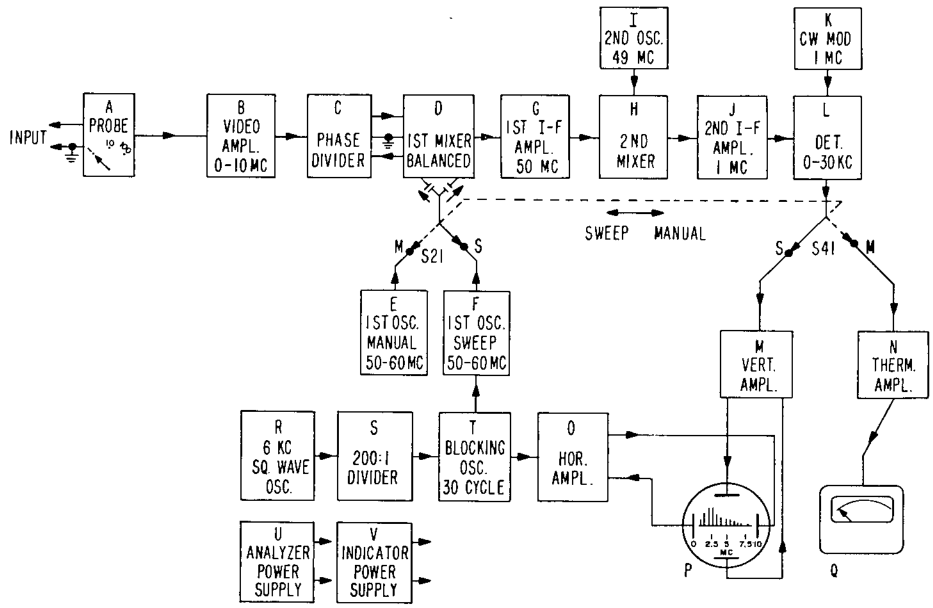


Fig. 2 - Video analyzer block diagram

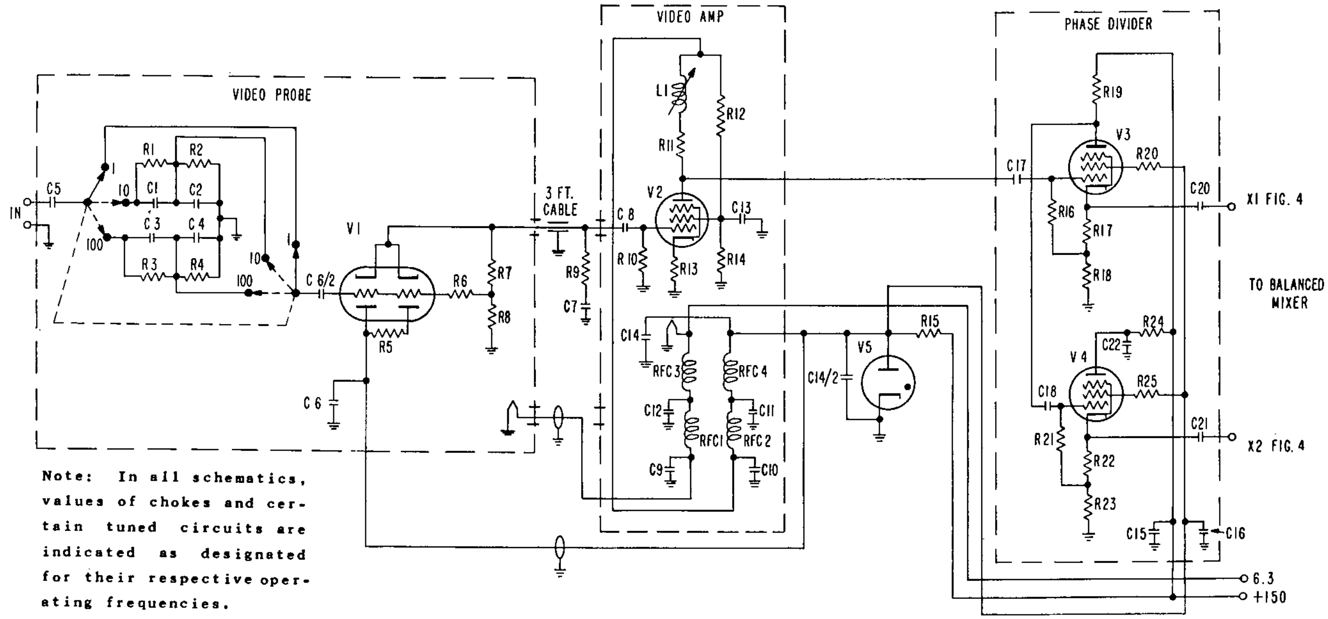
The operation of the instrument in the manual position follows the exact pattern in the visual presentation position, except that the local sweeping oscillator is replaced by a local, manually controlled oscillator, and the detected output is switched to an amplifier and associated thermistor bridge. The video frequency under observation is determined by the frequency of the local oscillator, which may be changed manually between 50 Mc and 60 Mc to cover the corresponding video band between 50 kc and 10 Mc. The meter, indicating the degree of unbalance (which is the measure of the energy in the second i-f amplifier envelope), may be calibrated using standards (as explained in Appendix A), and referred to in terms of absolute level by comparing the energy of the complex video signal under study to the corresponding energy of a sine wave.

DESIGN PROBLEM

As has been indicated, this type of analyzer is a double-conversion, superheterodyne receiver with the two special alterations mentioned. For the realization of a practical instrument, certain design characteristics require a certain amount of care in application. Beginning with the probe pickup, the main circuit elements of the analyzer will be described in the order followed by the video signal from the probe to the cathode-ray-tube indicator.

Probe and Preamplifier

The probe pickup unit (Fig. 3) is designed with a three-step attenuator preceding a cathode follower for use with high-level modulators, etc., having voltages exceeding the normal cathode-follower capabilities. A range of 20 db is provided, and this appears to be sufficient for most uses. The maximum voltage permissible on the high-range scale is 280 volts peak to peak. The frequency response of the cathode follower is flat within 0.5 db over the range of 50 kc to 10 Mc. The capacity loading is relatively light, and the probe can be brought physically in proximity to the voltage source whose characteristics it is desired to measure. The impedence of 15 $\mu\mu$f and 1 megohm resistance on the low range should have a minimum effect on the source whose spectrum it is desired to determine.



Note: In all schematics, values of chokes and certain tuned circuits are indicated as designated for their respective operating frequencies.

R1	900 K	R13	150 Ω	C1	5-12 μμf	C12	0.01 μf	V1	6J6
R2	100 K	R14	39 K	C2	33 μμf	C13	0.01 μf	V2	6AH6
R3	990 K	R16	220 K	C3	5-12 μμf	C14	0.01 μf	V3	6AH6
R4	10 K	R17	180 Ω	C4	330 μμf	C14/2	0.01 μf	V4	6AH6
R5	5.2 Ω	R18	820 Ω	C5	0.01 μf	C15	0.001 μf	V5	0B2
R6	2.7 M	R19	1.0 K	C6	0.01 μf	C16	0.01 μf		
R7	56 Ω	R20	120 Ω	C6/2	0.01 μf	C17	0.01 μf	L1	10 μh
R8	100 Ω	R21	220 K	C7	1.0 μf	C18	0.01 μf		
R9	56 Ω	R22	180 Ω	C8	0.01 μf	C19	0.01 μf	RFC1	50 Mc
R10	220 K	R23	820 Ω	C9	0.01 μf	C20	0.01 μf	RFC2	50 Mc
R11	1.2 K	R24	1.0 K	C10	0.01 μf	C21	0.01 μf	RFC3	50 Mc
R12	33 K	R25	120 Ω	C11	0.01 μf	C22	0.10 μf	RFC4	50 Mc

Fig. 3 - Video probe, amplifier, and phase divider

The video amplifier following the cathode-follower probe has a gain of 2.6 with a 0.5-db variation over a 10-Mc video band. This amplifier is inserted to restore the losses incurred in the cathode follower and the following phase divider.

Phase Divider

The video amplifier is followed by a pentode, cathode-follower type phase divider, which was found best over the 50-kc to 10-Mc band desired. The following balanced-mixer grids are fed from the low-impedance cathodes of the two 6AH6 tubes employed in the divider. The frequency response of the divider is essentially flat over the 50-kc to 10-Mc bandpass, and the over-all frequency response of the video chain from probe to mixer grids has a ±0.55-db variation.

Balanced Mixer

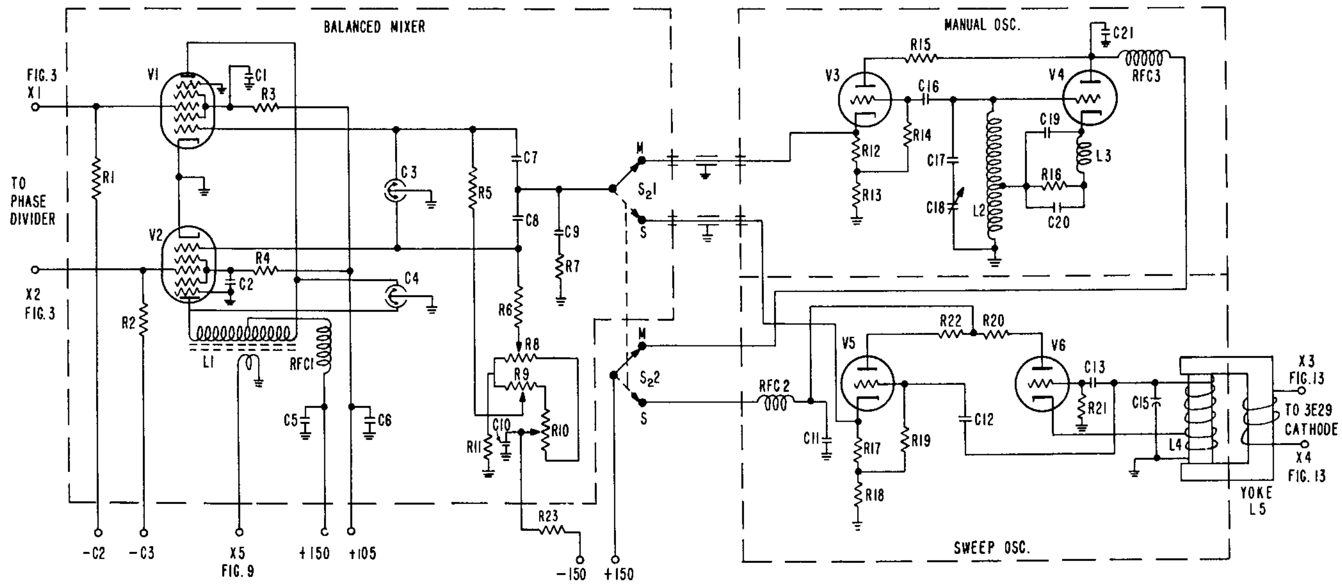
A balanced mixer (Fig. 4) is required to prevent overloading of the i-f amplifier. The local oscillator sweeps between 50 and 60 Mc, and when at the 50-Mc value, has the same frequency as the i-f amplifier center frequency. To keep the i-f amplifier from becoming saturated, a bridge circuit is utilized such that the local-oscillator voltage when applied in phase to the two mixer grids cancels out in the plate circuit. At the same time, the video signal is applied to the mixer out of phase and the mixed component passes freely to the i-f amplifier.

The balanced mixer employed is a more precisely constructed type than usually required with audio spectrum analyzers. This is required in part, because of lack of sufficient shielding within the 6BA7 tubes used in the mixer, which at 50 Mc, is not as effective as is desirable. Several other tubes tested were even less effective in this respect. A split-stator condenser with grounded rotor is bridged across the input and output of the mixer to phase the input and output properly, and thereby provide, as nearly as possible, a perfect null for the bridge circuit. Since the desired signal for the grid circuit of the first i-f amplifier is obtained from the lower side band only, it is necessary to eliminate any local-oscillator voltage from this point. This elimination is accomplished by a Faraday shield (Fig. 5) in the output transformer and the capacity-to-ground divider.

The bias potentials on the video-signal grids and on the local-oscillator grids are made variable in order that the gain of the two tubes may be equalized and thereby contribute to the more complete balancing out of the local-oscillator voltage and the symmetrical mixing of the video signal.

In balancing the mixer, it was found that a complete suppression of the local oscillator was not possible, but that its value could be made so small that the i-f amplifiers were not overloaded. In practice, the small spike at the left of the etched scale on the cathode-ray tube (which corresponds to zero video frequency and which occurs 30 times per second as the local oscillator sweeps across the first i-f), is of assistance in indicating whether the cathode-ray trace is correctly positioned and aligned.

The physical construction of the "squirrel cage" output transformer is shown in Fig. 5. The primary is shielded from the secondary by closely spaced copper wires parallel to the axis of the coil and bent away from the center of the coil to be attached to a circular nonshorting copper band that is in turn grounded. The secondary of the transformer is thus almost completely shielded from any stray capacity from the primary except for the desired magnetic coupling between the primary and the secondary. The primary is wound in a bifilar manner with one each of the alternate ends being connected together for the center tap. The two remaining ends go to the mixer plates.



R1	100 K	R12	130 Ω	C1	0.001 μf	C12	3.3 $\mu\mu f$	V1	6BA7	RFC1	50 Mc
R2	100 K	R13	750 Ω	C2	0.001 μf	C13	100 $\mu\mu f$	V2	6BA7	RFC2	50 Mc
R3	5 K	R14	330 K	C3	12-12 $\mu\mu f$	C15	1.5 $\mu\mu f$	V3	12AT7/2	RFC3	50 Mc
R4	5 K	R15	10 Ω	C4	12-12 $\mu\mu f$	C16	3.3 $\mu\mu f$	V4	12AT7/2		
R5	560 K	R16	2.2 K	C5	0.001 μf	C17	185 $\mu\mu f$	V5	12AT7/2		
R6	560 K	R17	130 Ω	C6	0.001 μf	C18	100 $\mu\mu f$	V6	12AT7/2		
R7	560 K	R18	750 Ω	C7	10 $\mu\mu f$	C19	75 $\mu\mu f$				
R8	50 K	R19	150 K	C8	10 $\mu\mu f$	C20	35 $\mu\mu f$	L1	0.7 μh		
R9	50 K	R20	10 Ω	C9	0.01 μf	C21	0.001 μf	L2	0.3 μh		
R10	50 K	R21	12 K	C10	0.1 μf			L3	0.44 μh		
R11	5 K	R22	5 Ω	C11	0.001 μf			L4	6 μh		
		R23	820 K					L5	(see text)		

Fig. 4 - Balanced mixer, sweep oscillator, and manual oscillator

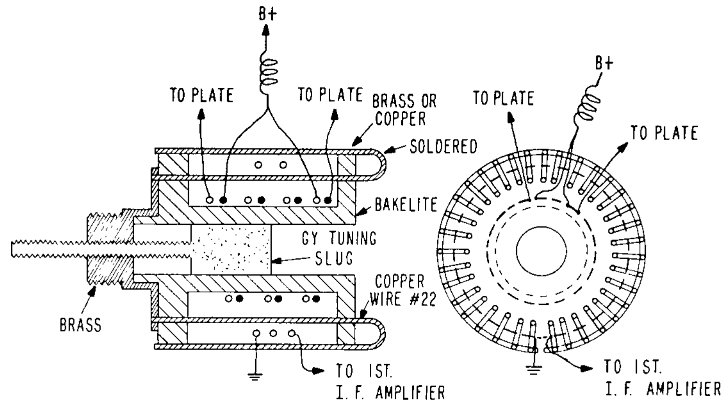


Fig. 5 - Mixer output transformer

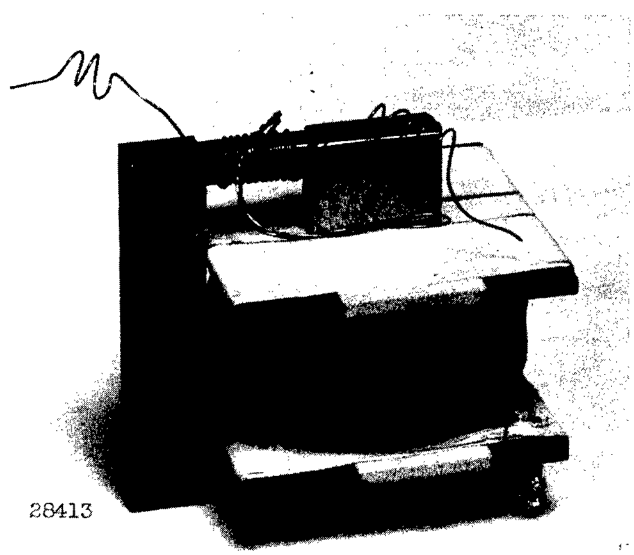


Fig. 6 - First local sweep oscillator

Local Oscillator (Sweep)

The sweep local oscillator employed is a compromise solution to the problem of designing a voltage source which could be rapidly changed in frequency, remain constant in amplitude, and be relatively free of harmonics. The design of the sweep circuit for the local oscillator was based on the principle that powdered iron, used as the core of the oscillator coil, suffers a reduction in permeability when magnetized. The powdered iron core of the oscillator coil* is mounted in the gap of a laminated iron yoke with the axis of the core parallel to the magnetic field (Fig. 6). When energized, the solenoid used to magnetize the yoke causes the frequency of the oscillator to increase. The yoke solenoid operates at a 30-cycle, linear, saw-tooth rate, being driven by a suitable blocking oscillator. By this means the local oscillator is swept linearly between 50 and 60 Mc.

*The powdered iron sample type Gy (Fig. 6) was 1/4 inch in diameter and 5/8 inch long; the yoke was wound with 55 ohms of No. 26 enameled wire.

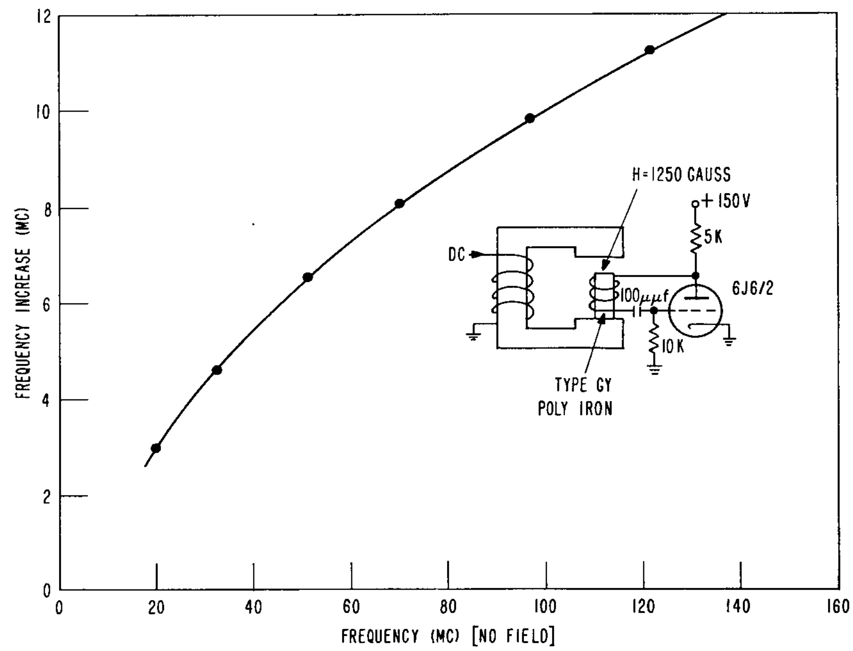


Fig. 7 - Frequency shift vs frequency of powdered iron at 1250 gauss field

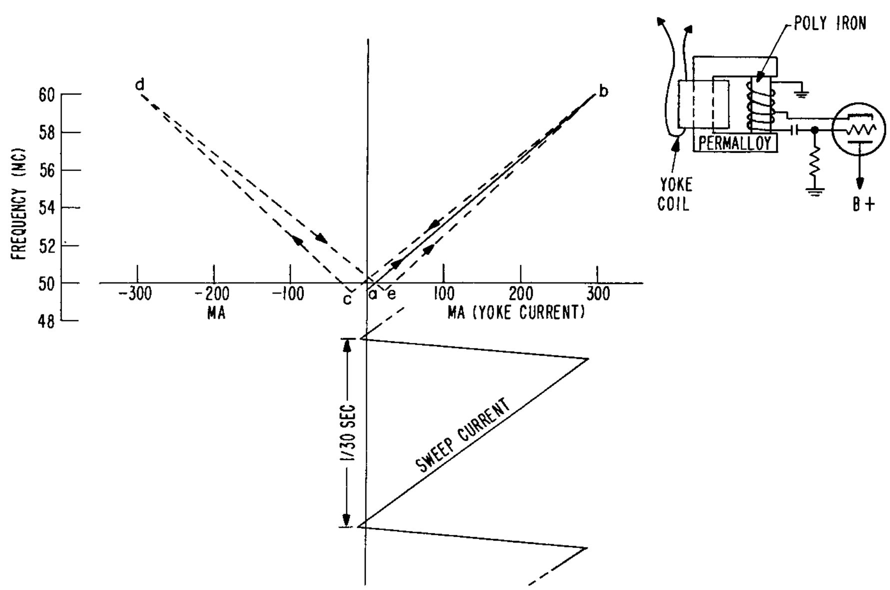


Fig. 8 - Frequency vs yoke current

Lack of time prohibited a careful study of the various iron powders available on the market today; a measurement of one of the few locally procurable at the time is shown in Fig. 7. The frequency change of the 50-Mc to 60-Mc oscillator (as used in this instrument) with respect to the solenoid current value (Fig. 8) indicates the linearity and hysteresis that may be expected. The hysteresis is not of importance because the return trace is blanked and not present on the scope. The linearity is sufficiently constant, so that the current in the yoke coil can be used to drive the horizontal deflection of the scope used for display.

In Fig. 8, it may be noted that the frequency change vs yoke current is approximately a linear function in one direction and for practical purposes the empirical numerical constants are

$$f = 50 + 35\ I \tag{4}$$

where f is the first local oscillator frequency (Mc) and I is the current in amperes in the yoke. A peak current of 0.28 ampere gave an oscillator frequency of 60 Mc, and a sweep time slightly less than 1/30 sec owing to the flyback time of the sweep return trace.

Local Oscillator (Manual)

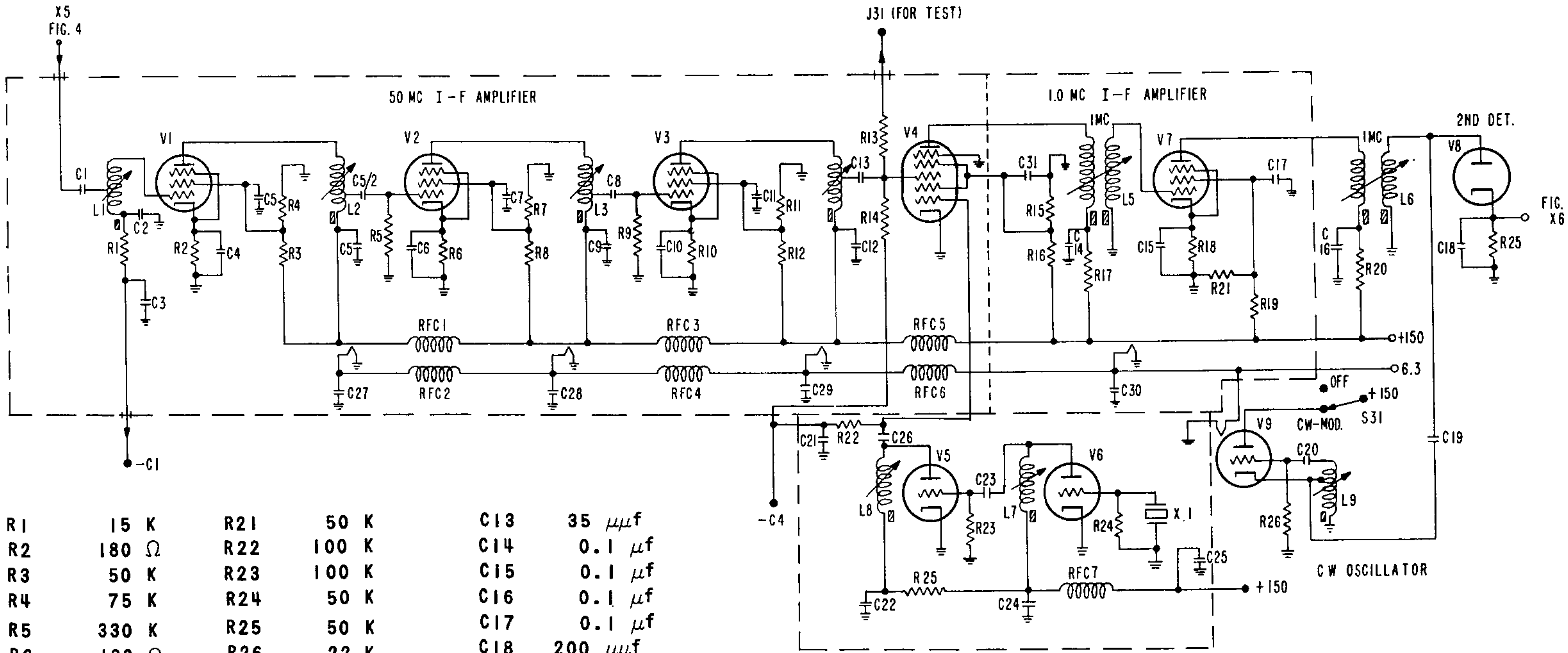
The analyzer provides, in addition to visual presentation, a manual position for operating with a power meter for reading directly the energy components of a complex wave form. The manual oscillator replaces the magnetically controlled sweep oscillator used in the visual display position. Likewise the power meter replaces the cathode-ray tube used in the visual position. Since the stability of the oscillator is required to be of a high order, some care is necessary in the design of this part of the circuit. The variable condenser should be sturdily constructed and should preferably have a straight-line-frequency characteristic to prevent excess bunching of the frequencies on one end of the dial. Figure 4 shows the oscillator circuit used.

First I-F Amplifier, Second Local Oscillator, Second I-F Amplifier, and CW Oscillator

Circuity for this section of the analyzer is shown in Fig. 9. The first i-f amplifier has a center frequency of 50 Mc and a bandwidth of 2 Mc. The bandwidth is determined by the Q of the coils used in the amplifier as well as by the loading of the tube capacities. At these frequencies, high-Q coils are difficult to manufacture; moreover, an over-all bandwidth of approximately 17.5 kc, which was desired in the instrument, could never have been obtained by using any coils manufactured today. Thus, a dual superheterodyne is employed with the lower i-f amplifier of a suitable bandwidth. The first i-f amplifier is single-tuned and has reasonable gain and average stability. A variable negative voltage is applied to the grid of the first tube to serve as a limited gain control.

The second local oscillator was chosen to operate below the first i-f (50 Mc) by one megacycle which is the second i-f value. No particular reason exists for the 1-Mc value except that transformers of this value were available. Since the difference frequency of 49 Mc is required to be extremely stable, a crystal second local oscillator was chosen, operating at 24.5 Mc and doubled to the desired frequency of 49 Mc, by V5 to insure a constant-frequency source.

The second i-f amplifier according to the previous discussion is required to be not less than 17.5 kc in bandwidth. The construction of the amplifier is of the double-tuned transformer type using slightly over-coupled elements. The gain is fixed, and is approximately 20 kc in bandwidth at the 3-db points, and approximately 30 kc at the 10-db points. Thus, the combination dual superheterodyne gives an effective i-f amplifier operating at a center frequency of 50 Mc with a final resolving bandwidth of 20 kc (3 db). Figure 10 shows the resolving bandwidth, with the dotted area giving the equivalent rectangular area as determined by a planimeter (Appendix A). This equivalent bandwidth as used for noise equivalent measurements is approximately 31 kc.



R1	15 K	R21	50 K
R2	180 Ω	R22	100 K
R3	50 K	R23	100 K
R4	75 K	R24	50 K
R5	330 K	R25	50 K
R6	180 Ω	R26	22 K
R7	75 K		
R8	50 K		
R9	330 K		
R10	180 Ω		
R11	75 K		
R12	50 K		
R13	3.9 K		
R14	100 K		
R15	50 K		
R16	50 K		
R17	5 K		
R18	200 Ω		
R19	50 K		
R20	5 K		

C1	0.001 μf	C13	35 μμf
C2	0.01 μf	C14	0.1 μf
C3	0.001 μf	C15	0.1 μf
C4	0.001 μf	C16	0.1 μf
C5	0.001 μf	C17	0.1 μf
C5/2	35 μμf	C18	200 μμf
C6	0.001 μf	C19	25 μμf
C7	0.001 μf	C20	350 μμf
C8	35 μμf	C21	0.001 μf
C9	0.001 μf	C22	0.001 μf
C10	0.001 μf	C23	27 μμf
C11	0.001 μf	C24	0.001 μf
C12	0.001 μf	C25	0.001 μf
		C26	25 μμf
		C27	0.001 μf
		C28	0.001 μf
		C29	0.001 μf
		C30	0.001 μf
		C31	0.1 μf

V1	6BA6	L1	1.5 μh	RFC1	50 Mc
V2	6BA6	L2	1.5 μh	RFC2	50 Mc
V3	6BA6	L3	1.5 μh	RFC3	50 Mc
V4	6BA7	L4	1.5 μh	RFC4	50 Mc
V5	12AT7	L5	1 Mc	RFC5	50 Mc
V6	12AT7	L6	1 Mc	RFC6	50 Mc
V7	6AH6	L7	24.5 Mc	RFC7	50 Mc
V8	1N38	L8	49 Mc		
V9	12AX7/2	L9	1 Mc		

X1 24.5 Mc Crystal

Fig. 9 - First and second i-f amplifiers, second local oscillator, and cw oscillator

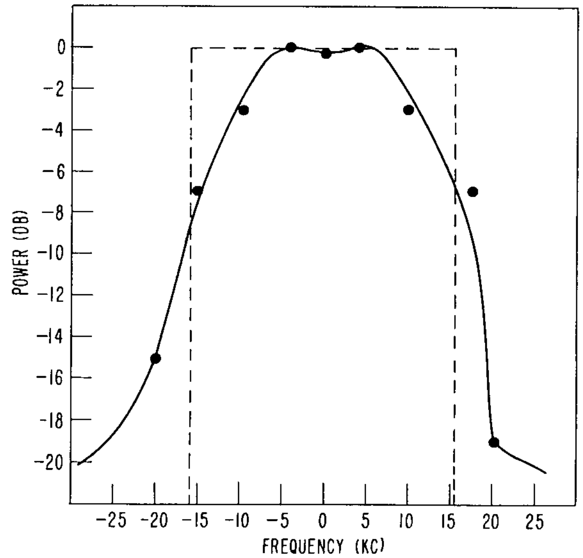


Fig. 10 - Bandpass of second i-f amplifier

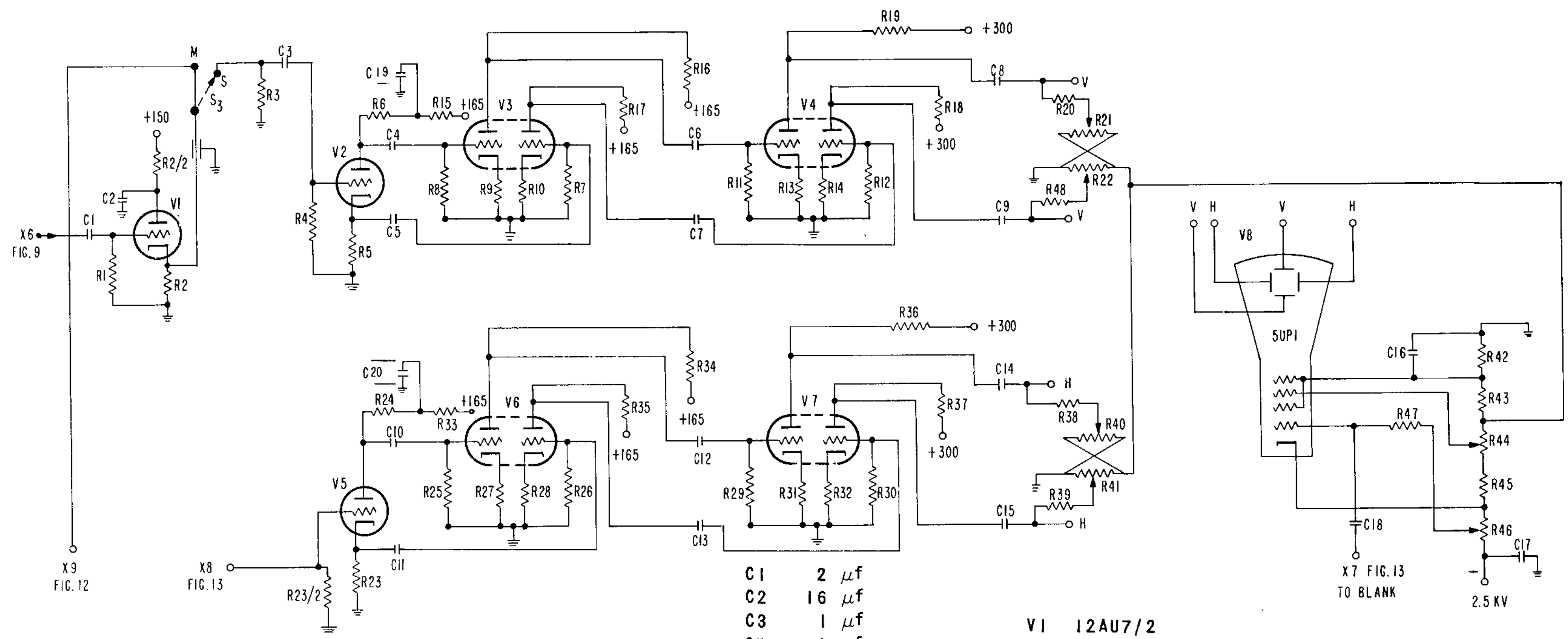
When measuring cw signals in the manual position no response occurs because the thermistor amplifier is not of the direct-coupled type. A 1-Mc cw signal is inserted at the final detector which beats with the signal coming from the second i-f amplifier, thereby restoring the ac component to be applied to the thermistor.

Visual and Manual Indicators

The indicator (Fig. 11) used for the visual display is a cathode-ray scope of relatively good low-frequency response because of the low sweep frequency used on the horizontal deflection (30 cycles). The high-frequency response extends well beyond the resolving power of the second i-f amplifier.

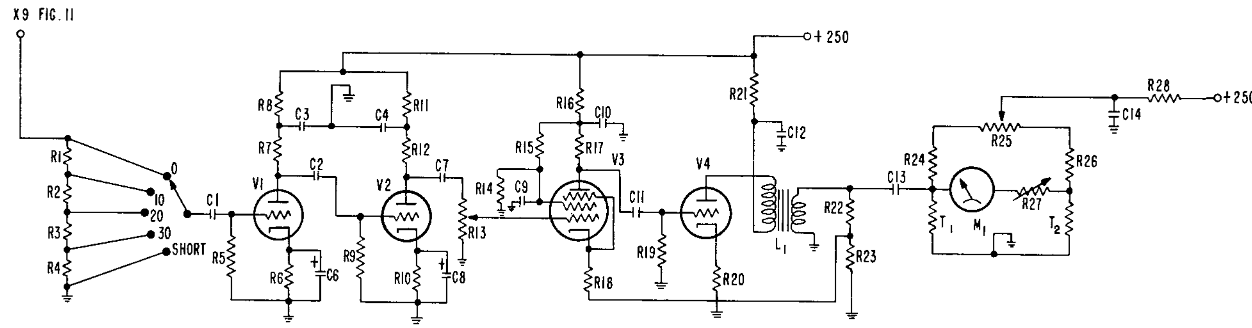
The indicator used to determine the relative response of the instrument when placed in the manual position (Fig. 12) consists of an amplifier whose output energy changes the resistance of a thermistor, causing a direct-current unbalance in the arm of a bridge which contains the thermistor as one arm. This method of power measurement is common in the audio-and radio-frequency spectrum. The meter employed as an indicator of bridge unbalance is a 0-50 microammeter reading directly in watts and dbm on a relatively long scale. The meter reading vs power was checked and the square-law response verified.

It should be observed that the two indicators are, respectively, linear and square-law with the latter giving a slower response suitable for manual operation since the thermistor has thermal lag. The use of the instrument in the manual position with reference to the comparison of random noise signals with their sine equivalent is explained in Appendix A.



R1	1 M	R17	5.1 K			C1	2 μf		
R2	4.7 K	R18	6 K	R33	3.9 K	C2	16 μf		
R2/2	4.7 K	R19	6 K	R34	18 K	C3	1 μf	V1	12AU7/2
R3	1.6 K	R20	1.2 M	R35	18 K	C4	1 μf	V2	12AU7/2
R4	1 M	R21	0.5 M	R36	18 K	C5	1 μf	V3	12AV7
R5	1.2 K	R22	0.5 M	R37	18 K	C6	0.5 μf	V4	5687
R6	1.2 K	R23	1.2 K	R38	1.2 M	C7	0.5 μf	V5	12AU7/2
R7	1 M	R23/2	0.5 M	R39	1.2 M	C8	1 μf	V6	12AV7
R8	1 M	R24	1.2 K	R40	0.5 M	C9	1 μf	V7	12AU7
R9	380 Ω	R25	1 M	R41	0.5 M	C10	1 μf	V8	5UP1
R10	380 Ω	R26	1 M	R42	2.5 M	C11	1 μf		
R11	1 M	R27	180 Ω	R43	2.5 M	C12	1 μf		
R12	1 M	R28	180 Ω	R44	2 M	C13	1 μf		
R13	680 Ω	R29	1 M	R45	1 M	C14	0.1 μf		
R14	680 Ω	R30	1 M	R46	0.5 M	C15	0.1 μf		
R15	3.9 K	R31	1.2 K	R47	0.5 M	C16	0.1 μf		
R16	5.1 K	R32	1.2 K	R48	1.2 M	C17	0.5 μf (3kv)		
						C18	0.01 μf (3kv)		
						C19	16 μf		
						C20	16 μf		

Fig. 11 - Visual indicator amplifier



R1	910 Ω	R16	33. K	C1	0.1 μf	V1	12AV7/2
R2	290 Ω	R17	50 K	C2	0.1 μf	V2	12AV7/2
R3	91 Ω	R18	560 Ω	C3	8 μf	V3	6AU6
R4	47 Ω	R19	500 K	C4	8 μf	V4	6C4
R5	500 K	R20	680 Ω	C6	50 μf		
R6	2.7 K	R21	3.9 K	C7	0.05 μf	L1	Plate to 200 Ω
R7	22 K	R22	390 Ω	C8	50 μf		
R8	56 K	R23	56 Ω	C9	16 μf		
R9	500 K	R24	130 Ω	C10	10 μf	T1	Thermistor
R10	2.7 K	R25	200 Ω	C11	0.1	T2	Thermistor
R11	56 K	R26	130 Ω	C12	20 μf	M_1	50 μ A
R12	22 K	R27	1 K	C13	4 μf		
R13	500 K	R28	60 K	C14	20 μf		
R14	180 K						
R15	160 K						

Fig. 12 - Thermistor amplifier

Sweep Generator

The method used to derive the 30-cycle sweep was chosen in order to facilitate certain experimental alterations in the analyzer planned for a later date. Otherwise simpler means would have been used. Rather than use the city mains 60-cycle frequency to drive a submultiple lock-over to produce a 30-cycle synchronizing signal, from which the 30-cycle blocking oscillator could be driven, a 6-kc square-wave signal was used to drive a 200:1 subdivider which was used as the 30-cycle synchronizing source. Figure 13 shows the circuit diagram. Once the synchronization signal is acquired (by whatever means), a monostable lock-over with appropriate time constants generates a 1800-microsecond pulse of a rather high voltage. This, in turn, is used to drive the 3E29 tube to cutoff condition, with the appropriate waveform to produce a linear current in the sweep-oscillator yoke solenoid. The principle applied resembles the techniques familiar to sweep problems in the television field. The same circuit used to obtain the blanking-oscillator negative pulse (V14 and V15) is duplicated exactly in V9 and V11 (except for altered time constants) to produce a negative pulse to blank the return trace of the cathode-ray tube, thereby avoiding any responses that might complicate the desired trace. A small voltage is developed across a 10-ohm resistor (R48) in series with the sweep solenoid; this voltage, when amplified, is used as the horizontal voltage for the visual-display indicator.

Power Supplies

Figure 14 gives a circuit diagram of the power supplies. Since the sweep circuit intermittently draws large currents at a 30-cycle rate, several regulators (both electronic and gas-tube types) are required. The principles have been known for a considerable time and should require only brief comment. The over-all physical make-up of the front and rear of the instrument is shown in Fig. 15 and 16.

EXAMPLES OF MEASUREMENT

The instrument described was designed primarily as a laboratory aid in the design of military equipment, particularly in noise measurements at high video frequencies. Figure 17 shows the spectra of a 6D4 gas triode. Figure 17a shows the noise spectrum of an unaltered 6D4 gas triode, while Fig. 17b indicates the same spectrum when corrected by appropriate filter nets to give a flatter frequency response. Figure 17c is the same spectrum as Fig. 17b except that the cw modulator switch is in the ON position. Note that the general shape of the spectrum envelope remains unchanged except that it is duplicated below the base line. Figure 17d gives the spectrum of a commercial noise generator (General Radio model 1390-A).

The limitations of the instrument for quantitative measurement of extremely short, low-duty-cycle pulse signals are recognized; however, some qualitative results are shown. Figure 18 shows the various spectra of a pulse voltage. Figure 18a indicates the shape of a one-microsecond pulse of a repetition rate of 5 kc. Figure 18b shows a spectrum of this one-microsecond pulse. The pulse spectrum should be the same if the pulse is of positive or negative polarity. Any unbalance of the first mixer is indicated by a difference in amplitude response. Figure 18c shows a 1/4-microsecond pulse spectrum.

The analysis of higher order duty cycle including square-wave (k = 0.5) signals is given for interest and possible application. If a square-wave generator (above 500 kc) is available, the instrument may be checked for fidelity of response by noting the relative amplitudes of the harmonics that are present. A study of Appendix B shows that the third harmonic of a square wave is 33% of the fundamental, and that the fifth harmonic is 20% of the fundamental.

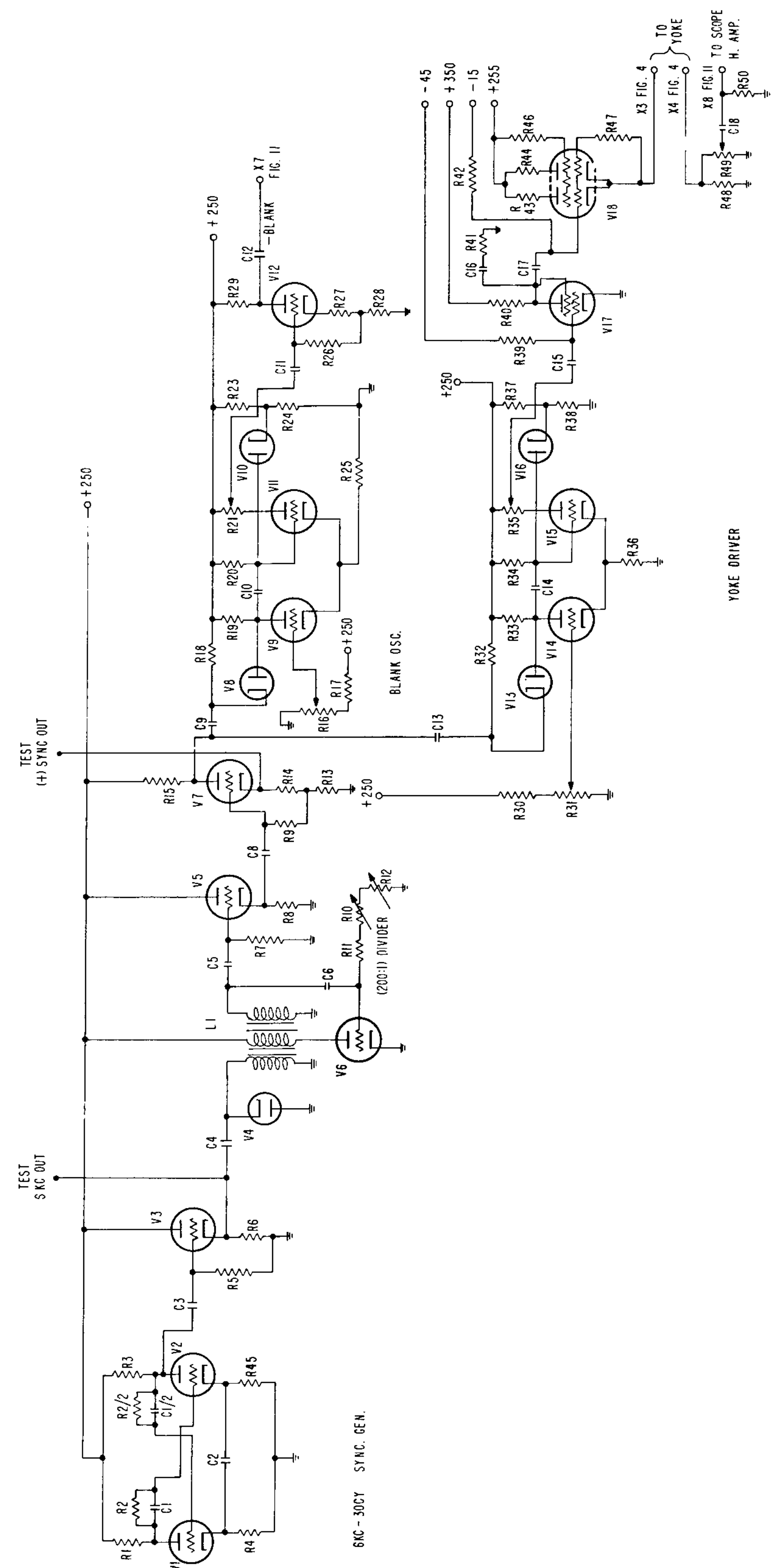


Fig. 13 - Sweep generator

COMPONENTS

R1	2.7 K	R26	1.5 M	C1	0.075 μf	V1	12AT7/2
R2	20 M	R27	1.5 K	C1/2	0.075 μf	V2	12AT7/2
R2/2	20 M	R28	6.8 K	C2	0.015 μf	V3	12AU7/2
R3	2.7 K	R29	8.2 K	C3	0.02 μf	V4	1N38
R4	100 K	R30	1.5 M	C4	150 μμf	V5	12AU7/2
R5	560 K	R31	50 K	C5	250 μμf	V6	12AU7/2
R6	2.7 K	R32	560 K	C6	0.005 μf	V7	12AU7/2
R7	100 K	R33	220 K	C8	0.001 μf	V8	6AL5/2
R8	2.2 K	R34	1.5 M	C9	50 μμf	V9	12AX7/2
R9	1.2 M	R35	250 K	C10	0.02 μf	V10	6AL5/2
R10	5 M	R36	10 K	C11	1 μf	V11	12AX7/2
R11	10 M	R37	270 K	C12	0.01 μf	V12	12AU7/2
R12	2 M	R38	27 K	C13	50 μμf	V13	6AL5/2
R13	6.8 K	R39	1.5 M	C14	0.01 μf	V14	12AX7/2
R14	1.5 K	R40	330 K	C15	1 μf	V15	12AX7/2
R15	8.2 K	R41	56 K	C16	0.07 μf	V16	6AL5/2
R16	50 K	R42	330 K	C17	4 μf	V17	6AQ5
R17	1.5 M	R43	15 Ω	C18	1 μf	V18	3E29
R18	560 K	R44	15 Ω				
R19	220 K	R45	100 K				
R20	1.5 M	R46	1 K			L1	145EW
R21	250 K	R47	1 M				
R23	270 K	R48	10 Ω				
R24	27 K	R49	100 Ω				
R25	10 K	R50	1 M				

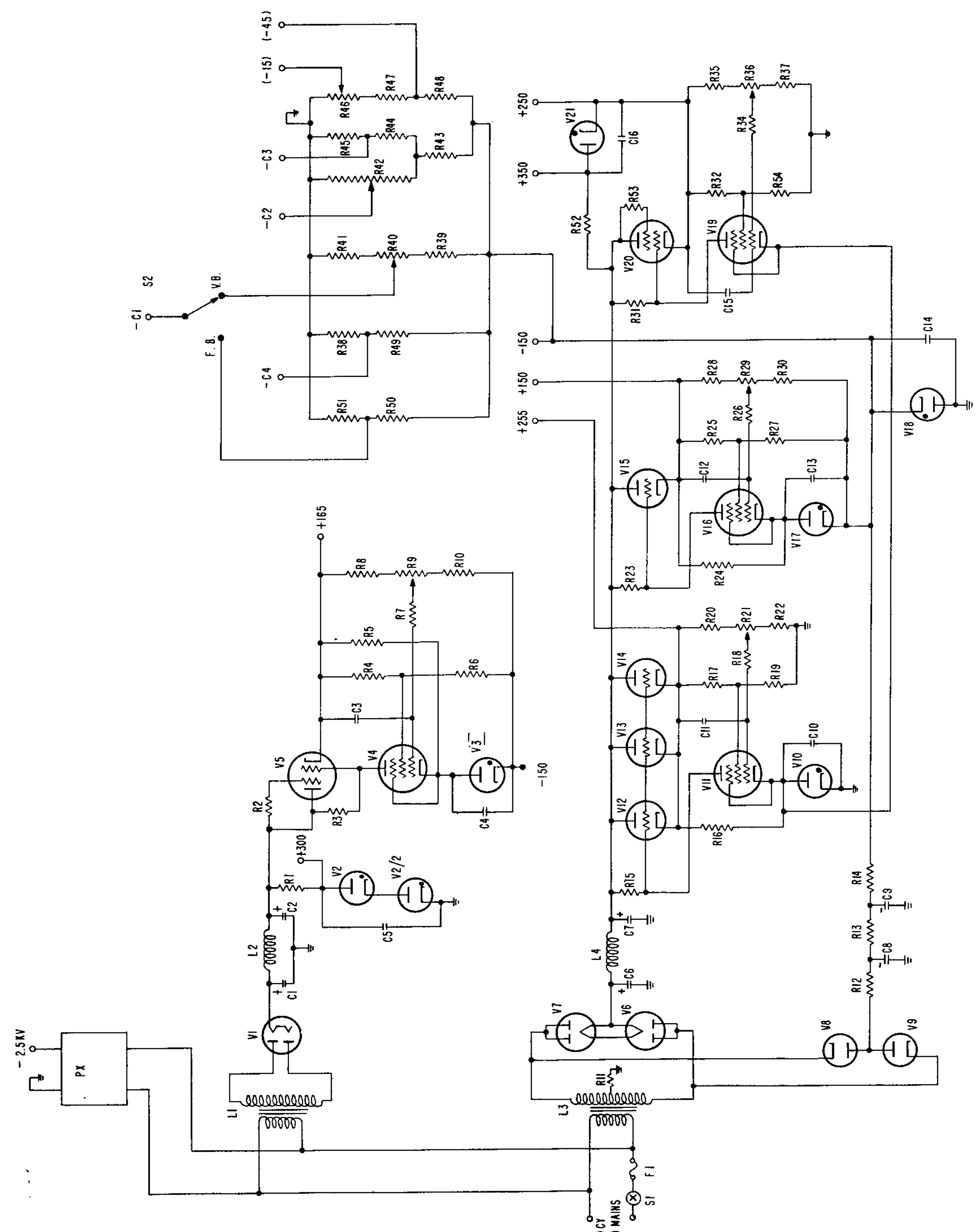


Fig. 14 - Power supplies

COMPONENTS

R1	1.5 K
R2	3.9 K
R3	2.7 M
R4	100 K
R5	82 K
R6	150 K
R7	220 K
R8	220 K
R9	50 K
R10	82 K
R11	1.0 Ω
R12	2.7 K
R13	7 K
R14	8 K
R15	2.2 M
R16	22 K
R17	220 K
R18	220 K
R19	100 K
R20	220 K
R21	25 K
R22	82 K
R23	2.2 M
R24	68 K
R25	220 K
R26	220 K
R27	150 K

R28	270 K
R29	25 K
R30	68 K
R31	1.5 M
R32	150 K
R33	150 K
R34	220 K
R35	270 K
R36	25 K
R37	100 K
R38	1.2 K
R39	150 K
R40	50 K
R41	3.9 K
R42	15 K
R43	68 K
R44	18 K
R45	18 K
R46	25 K
R47	27 K
R48	100 K
R49	150 K
R50	820 K
R51	20 K
R52	5 K
R53	100 Ω
R54	82 K

C1	20 μf
C2	40 μf
C3	0.1 μf
C4	0.01 μf
C5	0.01 μf
C6	4 μf
C7	250 μf
C8	8 μf
C9	8 μf
C10	0.01 μf
C11	0.1 μf
C12	0.1 μf
C13	0.01 μf
C14	0.01 μf
C15	0.1 μf
C16	0.01 μf

V1	5W4
V2	0A2
V2/2	0A2
V3	0B2
V4	6AK5
V5	6216
V6	5R4
V7	5R4
V8	6X4
V9	6X4
V10	VR75
V11	6AC7
V12	6AS7/2
V13	6AS7/2
V14	6AS7/2
V15	6AS7/2
V16	6SJ7
V17	VR75
V18	0A2
V19	6AU6
V20	6Y6
V21	0B2

L1	350-80 ma
L2	15H-85 ma
L3	400-340 ma
L4	10H-500 ma

F1	10A Fuse

PX	2.5 kv Packaged Supply

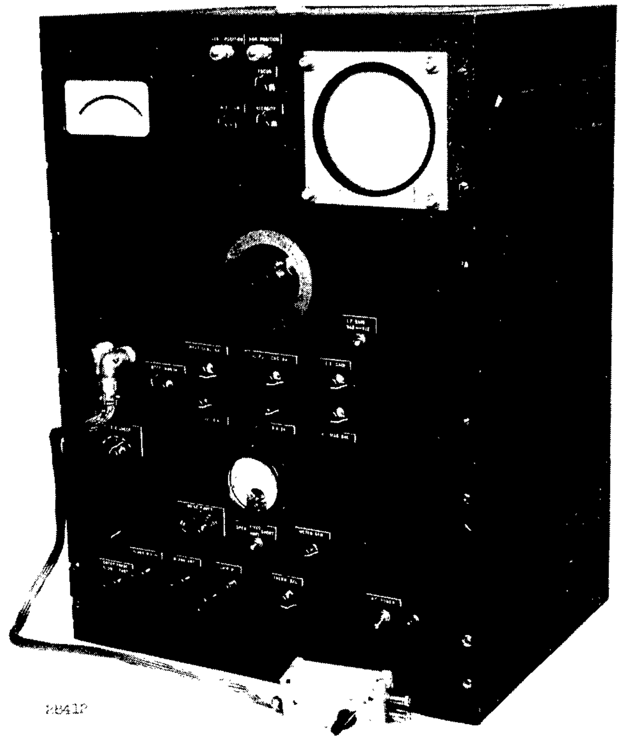


Fig. 15 - Front view of analyzer

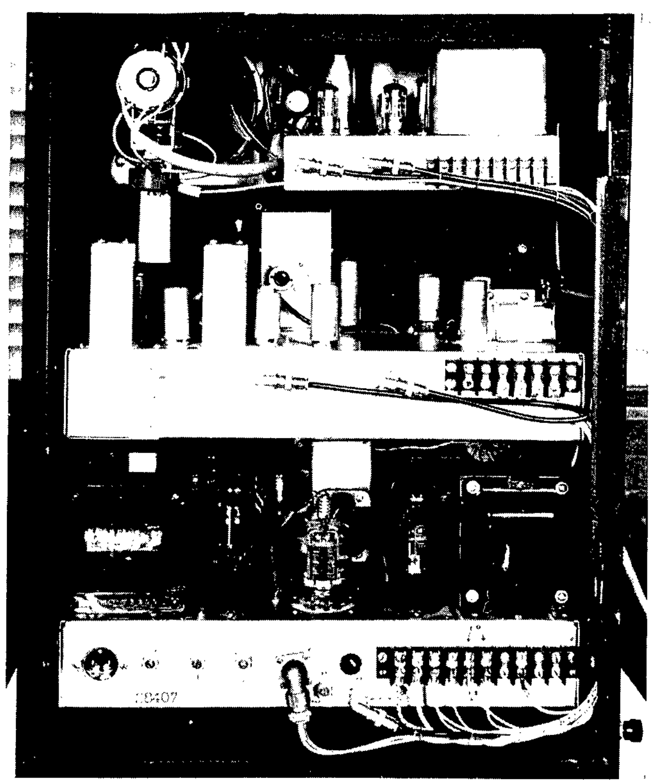


Fig. 16 - Rear view of analyzer

Figure 19 shows the spectrum of a square wave. The one-megacycle square wave from a commercial square-wave generator (Textronix model 105) is shown in Fig. 19a, and the corresponding spectrum is shown in Fig. 19b.

If the waveform symmetry of the square-wave generator is variable (Appendix B), the k factor may be varied so that the second harmonic (absent when $k = 0.5$), may be made equal to the third harmonic. The value found may be compared to the theoretical value, $k = 0.58$.

A brief analysis of the various factors that are used to calibrate the instrument, when in the manual position, is given in Appendix A. The noise equivalent (in terms of energy) as compared to sine-wave values, is shown with reference to the final resolving bandwidth and the response of the thermistor bridge included in the instrument. Some of the references given in RRL Report 411-260 * may be pursued for a fuller presentation than is given in the brief recapitulation included in Appendix A.

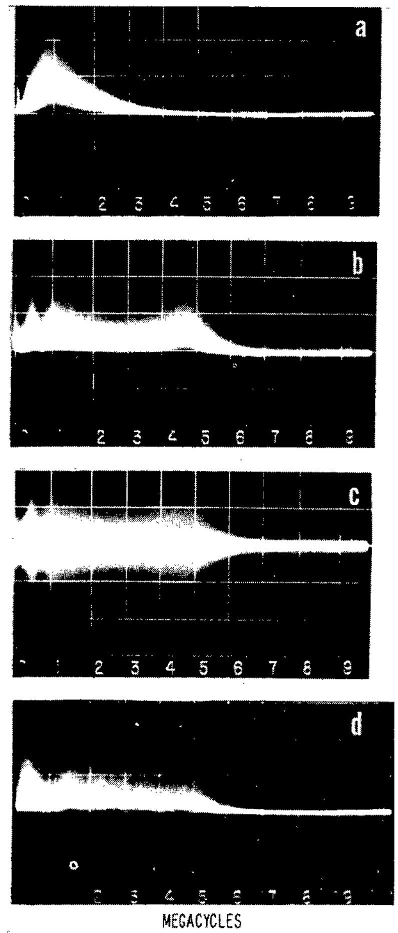


Fig. 17 - Spectra of a 6D4 noise triode. (a) Spectrum of a 6D4 noise tube, uncorrected; (b) spectrum of a 6D4 noise tube, partially corrected; (c) spectrum of a 6D4 noise tube, partially corrected, with cw modulation; (d) spectrum of a 6D4 noise generator (General Radio Model 1390-A).

CONCLUSION

The video analyzer described supplants several other methods of studying video noise spectra chiefly because a faster analysis is possible, in that the complete spectrum appears directly on a cathode-ray-tube screen.

Although this instrument is primarily designed to analyze high-level noise spectra, its sensitivity is sufficient to examine a normal video detection output directly without amplification, and at the same time is capable of examining the output of most medium- and high-power video amplifiers and modulators.

The instrument requires a minimum of operating skill and maintenance, and during approximately two years of experimental service has performed reliably.

With some reengineering, a suitable model could be designed for service use.

The size of the instrument could be considerably reduced without sacrificing the main operating features.

*Jastrum, P. S., "The Circuit Theory of Noise," Harvard University RRL Report 411-260, November 1, 1945

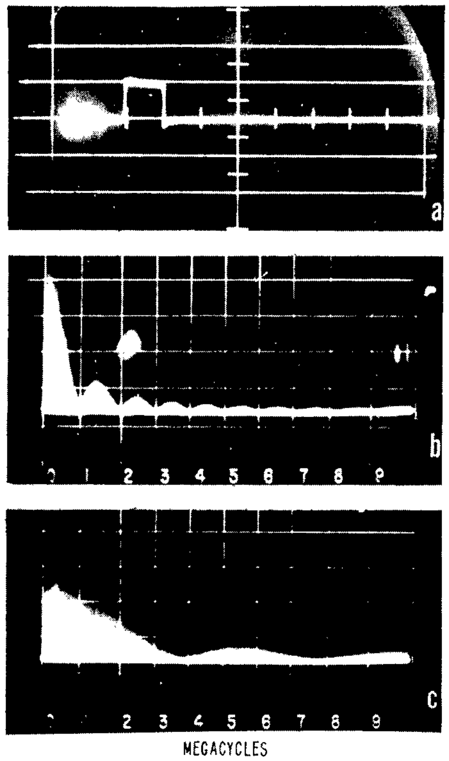


Fig. 18 - Spectra of pulses. (a) one-microsecond pulse; (b) spectrum of a one-microsecond pulse; (c) spectrum of a 1/4-microsecond pulse.

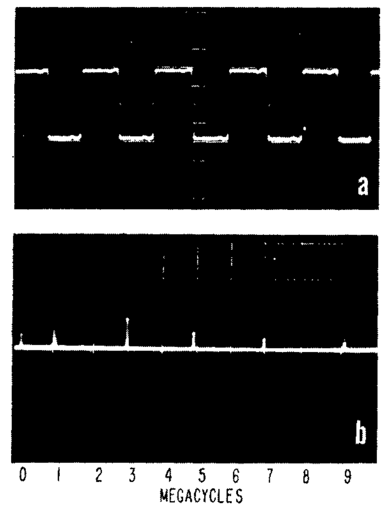


Fig. 19 - Spectrum of a square wave. (a) one-megacycle square wave (Textronix Model 105); (b) spectrum of a one-megacycle square wave.

ACKNOWLEDGMENTS

Many suggestions have been given the writer regarding the characteristics desirable in an analyzer. Mr. W. E. Withrow's special interest in suggesting, among others, the idea of including a manual position and a scaled meter, is acknowledged.

* * *

APPENDIX A
Application of Analyzer in Study of Noise

Many references to the problem of random-noise power measurements with a thermistor indicator have appeared in the literature from time to time. The brief recapitulation given here uses nomenclature adapted from a Radio Research Laboratory report,* although perhaps other treatments would be as acceptable.

The mean-square value of a periodic function is given as

$$V(t) = \sum_{n=0}^{N} A_n \sin n \omega_0 t$$

or

$$\overline{V}^2 = \frac{1}{2} \sum_{1}^{N} A_n{}^2 \tag{A1}$$

The latter equation states that the mean-square value of the time function integrated over a number of periods is equivalent to one-half the sum of the squares of the amplitudes of the components of the spectrum.

If 1/2 $A_n{}^2$ is replaced by $y^2(f_n)\ \Delta f$ and the limits taken, we have the differential form

$$\frac{1}{2} A_n{}^2 = y^2(f_n)\ \Delta f;$$

$$\overline{V}^2 = \sum_{n=0}^{N} y^2(f_n)\ \Delta f$$

or

$$\overline{V}^2 = \int_0^{\infty} y^2(f)\ df \tag{A2}$$

The familiar expression results from the theoretical treatment backed by the following experimental procedure. If, for simplicity we rule out discrete frequency periodic voltages in an otherwise continuous spectrum, proceed to measure the spectrum of a given signal, plot all the individual power levels on a curve, and sum up the power under this plot, it should equal the power of all the combined frequency elements of the signal when measured by some nonfrequency-conscious, heat-measuring device. This checks out fairly well experimentally, supporting the procedure outlined above.

By the imaginary process of having a variable sampling filter which may be freely moved, at will, over the spectrum in question, together with the provision that the bandpass of the filter may be reduced to a differentially small quantity, it may be deduced by theoretical reasoning that the mean-square value of the filter output is proportional at very

*Jastrum, P. S., "The Circuit Theory of Noise," Harvard University RRL Report 411-260, November 1, 1945 narrow bandwidths, to the bandwidth of the filter. By assuming that the noise over the bandpass is constant, it may be deduced further that the output of the filter gives a response which is proportional to the power of the spectral components under the bandpass, but which must be determined experimentally for each filter and indicator. Although the comparison between the rms value of noise and a sine wave is somewhat meaningless when the bandpass of the analyzing net approaches zero, it goes without saying that all measuring devices have a finite bandwidth and that the two signals (sine and noise) may be compared in practice. When the various corrections necessitated by the filter characteristics are made, it may be stated that the output amplitude may be represented as

$$V = P(t) \sin \omega_0 t \tag{A3}$$

which may be designated as a sine wave, the amplitude of which varies in a random manner; the center frequency is $\omega_0/2\pi$ and the envelope varies from zero to one-half the bandwidth of the filter, Δf.

In the above discussion for the value of the power relation (Eq. A2) the total band is indicated, and if we are to define the relation for a narrow band within the broad spectrum, the following relation is the narrow-band equivalent

$$y^2(f) = \lim_{\Delta f \to 0} \frac{E^2(f, \Delta f)}{\Delta f} \tag{A4}$$

where $E \approx$ filter output rms.

Obviously no instrument can have a bandwidth of zero. The numerator of Eq. (A4) is the power equivalent of the noise spectrum within the Δf band, and the rms value may be written in volts as

$$y(f) = \frac{E(f, \Delta f)}{\sqrt{\Delta f}} \tag{A5}$$

The noise rms value is given in volts per square root cycle, kilocycle, or megacycle.

To calibrate a given instrument, a calibrated sine wave is introduced into the instrument, the manual local search oscillator is tuned to the frequency, and the gain is adjusted to zero db. The sine-wave reference generator is now varied over a given number of kilocycles until the gain is 20 db down. The power under the curve is integrated with a planimeter and a hypothetical rectangle constructed with the area equivalent to the area under the curve drawn above. This bandwidth is the equivalent (over 95%) required to give a symbolic rectangle that has the height of the maximum sine-wave response and the area equal to the actual bandpass (Fig. 9). In using the analyzer, the noise spectrum is searched for the highest amplitude point and the adjustments left fixed; then a sine wave is introduced and the voltage required to give the same response as the noise spectrum is noted.

If

$$E_{cw}^2 \approx E^2(f, \Delta f) \tag{A6}$$

$$y(f) \approx \frac{E_{cw}}{\sqrt{\Delta f}} \tag{A7}$$

where Δf = 31 kc.

When the peak noise value has been computed (in volts per root cycle; kc or Mc), the other parts of the spectrum may be measured as so many db down from the peak value,* using the db scale of the thermistor meter. The procedure outlined here is believed to be more accurate than the methods adopted by many analyzers, in that the gain of the various i-f amplifiers (normally fixed) is far from constant because of age and gain change due to ambient temperature rise of the equipment. The probability of errors of measurement due to the possibility of bandwidth change is more remote than that due to the change in gain; and since a direct reference to a standard signal generator requires the gain to be maintained only over the short time that the measurement is being made, the error of the standard is the principal one.

The maximum sensitivity of the instrument is ascertained on the basis of the rms voltage required to produce an output 3 db above the inherent circuit noise of the instrument. A value of 880 microvolts rms in the visual position and 38 microvolts rms in the manual position was found. Using the manual sensitivity as a reference, the voltage per root cycle may be computed to be 0.216 microvolt per root cycle.

---

*Reference is made to the manner that data has been presented.

(a) Volts per root cycle (db above one microvolt):

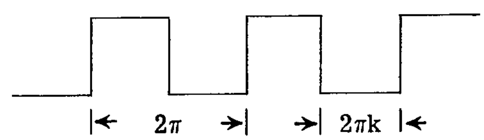
$$\frac{20 \log \text{volts (cw) rms}}{\sqrt{31{,}000} \times 10^{-6}} = 20 \log \left[\text{volts (cw)} \times 5.7 \times 10^{3}\right]$$

(b) Volts per root kilocycle (db above 10 microvolts):

$$\frac{20 \log \text{volts (cw) rms}}{\sqrt{31} \times 10 \times 10^{-6}} = 20 \log \left[\text{volts (cw)} \times 18 \times 10^{3}\right]$$

* * *

APPENDIX B
Application of Analyzer to Rectangular Waves



The general form for the Fourier series for a rectangular wave may be written

$$Y = V \left\{ k \frac{2}{\pi} \left(\sin k\pi \cos x + \frac{1}{2} \sin 2k\pi \cos 2x \cdots \frac{1}{n} \sin nk\pi \cos nx \right) \right\} \tag{B1}$$

which reduces to the following for a square wave (k = 0.5)

$$Y = V \left\{ 0.5 \frac{2}{\pi} \left(\cos x + \frac{1}{3} \cos 3x + \frac{1}{5} \cos 5x \cdots \frac{1}{n} \cos nx \right)_{n\,=\,\text{odd}} \right\} \tag{B2}$$

The even harmonics are cancelled and the odd are related to the fundamental as indicated in the ratio of 1:0.33:0.2: · · · etc., and may be shown on the indicator (Fig. 19b).

If the value of k is altered, the second and the third harmonic may be made equal, and the value of k may be compared to the theoretical value of 0.58.

From the general form above

$$\frac{1}{2} \sin 2k\pi \cos 2x - \frac{1}{3} \sin 3k\pi \cos 3x = 0 \tag{B3}$$

If x is made zero

$$\frac{1}{2} \sin 2k\pi - \frac{1}{3} \sin 3k\pi = 0$$

$$\frac{3}{2} = \frac{\sin 3k\pi}{\sin 2k\pi} = \frac{3 \sin k\pi - 4 \sin^3 k\pi}{2 \sin k\pi \cos k\pi} = \frac{3 - 4 \sin^2 k\pi}{2 \cos k\pi}$$

$$4 \cos^2 k\pi - 3 \cos k\pi - 1 = 0 \tag{B4}$$

$$k\pi = 14.5^0 \ (+\ 90^0, \text{ or } 104.5^0) \text{ second quadrant}$$

$$\frac{104.5^0}{360} \times 2\pi = 1.82$$
$$k\pi = 1.82$$
$$k = 0.58$$

The value of k is determined by an auxiliary oscillograph that has sufficient bandwidth to present correctly the rectangular wave components.

* * *

NAVY-DPPO PRNC. WASH., D.C