The 1948 National Convention of the IRE: held at New York City, 22-25 March

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RADIO AND ELECTRICAL ENGINEERING DIVISION

THE 1948 NATIONAL CONVENTION OF THE IRE

HELD AT NEW YORK CITY, 22-25 MARCH

OTTAWA

APRIL, 1948

THE 1948 NATIONAL CONVENTION OF THE INSTITUTE OF RADIO ENGINEERS
NEW YORK CITY
MARCH 22nd-25th

Introduction:

A few introductory remarks seem indicated, partly by way of apology for an incomplete report on the Convention. Four delegates from the Division were originally selected to attend the Convention but one was forced to drop out at the last minute. Even with four representatives it would not have been possible to cover all the events satisfactorily. However, it turned out that a good many of the papers were really not worth covering, although it was rather difficult to assess their quality in advance. Some of the papers had already appeared, and others were re-hashes of old stuff. Attempts were made to move from one lecture room to another in order to take in specific papers but these attempts were frequently thwarted, sometimes because the session chairman changed the order of the papers, but more frequently because of the terrific congestion. We were told there were over fifteen thousand registered by the second day and as a result even the largest lecture halls had standing room only during the more popular papers. Thus, we either strived to get a good seat before the session started and stayed there, or we wriggled in and stood at the back, usually in the dark, without being able to make any notes. A few papers were exceptionally good and well worth the trip.

On the whole, though, and despite the pessimistic remarks above, the Convention was of great interest to those attending. The greatest benefits to be obtained, of course, are derived from the new contacts one makes and the renewal of old acquaintances. This is largely a personal matter and is admittedly of little value to the person staying at home and reading this somewhat sketchy account. For that reason it is suggested that as many persons as possible should attend the main I.R.E. Convention each year and that this privilege should be rotated throughout the scientific staff. More information and more new ideas are accumulated over a cup of coffee or an evening of rag-chewing with contemporary engineers and scientists than can be gleaned from attendance at the sessions.

D.W.R. McKinley
C.W. McLeish
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LECTURES ATTENDED BY MR. MCLEISH

A. TECHNICAL SESSION ON NETWORKS

Paper #6. "Properties of Some Wide-band Phase-splitting Network", by D.G.C. Luck, RCA.

Dr. Luck discussed some kinds of well-known branch circuits which provide constant amplitude response with respect to frequency. Two of these types are shown in figs. 1(a) and 1(b). Both are frequency sensitive as far as phase shift is concerned. A third circuit, as shown in fig. 1(c), can have a fairly broad band characteristic. The phase characteristic of each of the shunt paths is shown in fig. 1(d). Frequencies f_1 and f_2, where ϕ is zero. The output phase shift is taken as that involved in the difference voltage $(e_1 - e_2)$, which is $(\phi_1 - \phi_2) = x$. The value of x can be held fairly constant over a band of frequencies controlled by the Q and the frequency separation of the two parallel circuits.

As an example, Dr. Luck quoted a figure of phase deviation of 3/4° for a nominal shift of 10° over a frequency range of one octave. Fig. 1(e) shows typical phase response curves.

Reference: R.V. Dome Dec. '46 "Electronics".

Paper #7. "Theory and Design of Constant-Current Networks", presented by C.S. Roys, Syracuse University.

Mr. Roys reviewed series resonant circuits of which descriptions are to be found in papers by Steinmetz, Green and Cooper. In general, these are sensitive to load power factor.

A circuit, independent of load power factor, termed the monocyclic square, was described. Fig. 2(a) shows this network which has the main disadvantage that under optimum conditions only one-half the generator volt amperes are available in the load. When the load Z_o is $\frac{1}{2} X_L$ and $X_L = X_C$, we get the best efficiency. Current distortion is zero when the load is resistive and equal to $\frac{1}{2} X_L$.

Paper #8. "New Parameter Adjustment Method for Network Transients", by M.J. DiToro, Federal Telecommunications Labs. Inc.

A paper by the author, published in Proc. IRE for January, 1948, was referred to by Dr. DiToro. A specific application to video circuit design was described, showing how the parameters may be adjusted to provide a much faster rise time than that obtained with critical damping and yet produce negligible overshoot (for some purposes).

Response curves were shown comparing the transient response of circuits designed for (a) critical damping (by shunt compensation), (b) maximum amplitude response (by series compensation), and (c) optimum transient response (by the new parameter adjustment method).

The design procedure seemed particularly applicable to television video circuits where a small overshoot is allowable if rise time can be correspondingly reduced.

<u>Paper #9</u>. "Application of Tehebyschef Polynomials to the Design of Band-pass Filters", by M. Dishal, Federal Telecommunications Labs.

In spite of its formidable title this paper described a fairly practical means of predicting mathematically the circuit parameters required in "N" cascaded tuned circuits to provide certain "nose" and "skirt" bandwidths.

The expression for amplitude response of N cascaded tuned circuits is expanded into a series, known mathematically as the Tehebyschef polynomial. Knowing the required bandwidths at two levels of attenuation it is possible to solve for the coefficients of the polynomial and derive therefrom the values of Q for the tuned circuits. The method is only exact when the bandwidth is small compared with carrier frequency. For large bandwidths an extension of the analysis has apparently been done using log f as the frequency base instead of f.

<u>Paper #10</u>. "A Simplified Negative-Resistance Q Multiplier", by H.E. Harris of MIT.

A practical method of increasing the effective Q of a tuned circuit was described using the basic circuit shown in fig. 3(a). Letting K = no-feedback gain of the amplifier, the condition of stability (i.e. no oscillation) for a given charge of gain ΔK was given as

$$\frac{\Delta K}{K} = \frac{1}{\frac{Q_{eff}/Q_o}{K/2} - 1}$$

For maximum stability with respect to G_m changes in the cathode-follower tube, $C_1 = C_2$. An example was cited of a coil whose normal Q was 112 but with the use of the circuit the effective Q was raised to 25,000. Q multiplications of 10 times are considered quite practicable.

B. TECHNICAL SESSION ON AMPLIFIERS

Paper #27. "Low Noise Amplifier", by Wallman, Macnee, and Gadsden of MIT.

Discussion was of course confined to the input stages of an amplifier. Here the use of triodes was recommended because of their inherently lower shot noise for a given G_m.

A circuit, shown in fig. 4(a), employing two triodes in cascade has the very low noise figure of 1.35 at 30 mc per carrier frequency and .6 mc bandwidth. This circuit is superior to that using a cathode-coupled input tube because the latter has a gain equivalent to a tube having only one-half the G_m of the first triode (due to input loading) whereas the circuit described has an overall gain equal to $G_{m1}R_2$. ~~Tubes used~~ were type 6AK5. The first tube was neutralized but the value of L_n was not found to be critical in this circuit over the bandwidths used in practice.

Paper #28. "Phase Distortion in Audio Systems" - L.A. DeRosa - Federal Telecommunications Labs.

This paper discussed the effects on the ear of phase distortion and transients. The ear is said to behave as a transient analyzer during large amplitude changes and is extremely sensitive to "attack rate". After the transient is over, the ear becomes sensitive to harmonic content. Phase distortion has a marked effect on the response of the ear, particularly in so-called "high-fidelity" systems where a large frequency range is reproduced.

Reference: Jour. Acc. Soc., July 1947, paper by Stevens, Davis and Lurie.

Paper #29. "Visual Analysis of A.F. Transient Phenomena" by D.E. Maxwell, C.B.S.

Pulsed audio frequency energy is impressed on the audio system to be tested and the response is examined on a cathode ray tube. This was shown to be a convenient and practical means of testing audio amplifiers, speakers, and studios. Testing of speakers is done by applying the pulse to the voice coil, and using a microphone at close range as the pickup unit. Echoes in the room do not interfere with the analysis if the microphone is much closer to the speaker than any reflecting surface. Studios are measured for reverberation time by means of a speaker and microphone arrangement, the characteristics of which are already known.

Paper #30. "Square-Wave Analysis of Compensated Amplifiers"
P.M. Seal, University of Maine.

This paper dealt solely with low-frequency compensation. The response of the amplifier to square-wave input was examined in terms of percent variation of amplitude along the top of a "square wave". Referring to a circuit diagram using a conventional method of low frequency compensation (fig. 5(a)), Mr. Seal showed that an output wave, having the same amplitude at the end as at the start, was obtained if $R_F = 10R_o$, and $R_F C_F = R_g C_c$.

A slight hump was produced in the middle but the overall deviation from the input square wave was considered about at a minimum. The response of such a circuit is good down to a frequency given by $\frac{1}{2\pi R_g C_o}$.

Paper #31. "A New Figure of Merit for the Transient Response of Video Amplifiers" - presented by L. Mautner - Dumont Labs.

Overshoot is defined in this paper as the maximum deviation, in terms of amplitude, of the circuit to a step function. Rise time is defined as the 10% to 90% time of rise. The figure of merit proposed is

$$F = 1000T^{-1}\epsilon^{-100\gamma^2} \quad \text{where}$$
$$\gamma = \text{overshoot}$$
$$T = \text{rise time}$$

The factor of 1000 raises F to a convenient figure, and the factor 100 in the exponential limits the permissible overshoot rate change in F. appears to the second power to make F independent of the sign of .

Several networks were evaluated in terms of the new figure of merit, which it is understood would have greatest application to television design. Figs. 6(a) and (b) show two of the networks.

Paper #32. "Distributed Amplification",
presented by E.L. Ginzton, Stanford University.

This paper described briefly a very interesting amplifier in which frequency response is limited at the high frequencies only by grid loading effects of the tubes used. Cascading is no limitation on bandwidth. Fig. 7(a) shows the basic circuit of the distributed line amplifier. It was disclosed that the optimum voltage gain of a stage for any bandwidth is the natural logarithm base $\epsilon = 2.718$. Additional voltage gain is obtained by cascading stages. The number of tubes in a stage is determined by the bandwidth. To take care of line losses at high frequencies, some plates of a stage are paired as shown in fig. 7(b).

The signal noise ratio of the amplifier is proportional to the number of tubes per stage, and noise due to induced grid noise is proportional to the (number of tubes per stage)^3, so this factor limits the use of some tubes.

As an example of the application of the type of amplifier, a response curve was shown of an amplifier having 2 cascaded stages of 7 tubes each which had quite a uniform response from zero to 150 mc/sec with a gain of 18 db.

C. TECHNICAL SESSION ON TRANSMISSION

Paper #54. "Optimum Geometry for Ridged Waveguide", presented by W.E. Waller, Polytechnic Research and Development Co., Brooklyn, N.Y.

The optimum proportions of a ridged guide as shown in fig. 8(a) were discussed. Considerations are the power handling capacity and the attenuation. The properties suggested are s/a = 1/4 and d/b = .2. A ridged guide which operates successfully over the band 4 - 10 Kmcs. As the following dimensions:

a = 1.196"
b = .598"
d = .201"
s = .359"

For the guide the attenuation per foot is about .35 db. At high frequencies this type of guide has better power handling capacity and lower attenuation than coaxial line, and has a broad band characteristic compared to rectangular guide.

Paper #55. "Fields in Non-metallic Waveguides", R.N. Whitmer, Rensselaer Polytechnic Institute.

For a cylindrical dielectric waveguide, propagating the mode next to cut-off frequency, the field at the surface is high - about 40% of the field at the center. However, the attenuation of this field outside the guide is also high - about 30 db per radius.

Attempts were made to reduce the external field by making the guide dielectric constant at any point in the guide roughly proportional to the universe of the radius at this point. This reduced the surface field considerably but also decreased the attenuation of the field outside the guide so that the net result was no improvement over a uniform dielectric.

Advantages of dielectric guide appear chiefly on very short wavelengths when manufacturing costs on metal waveguide become high due to the close tolerances required.

Paper #56. "A Wide-band Waveguide Filter Structure", by S.B. Cohn - Harvard University.

This paper was a theoretical analysis of a waveguide filter as shown in fig. 9(a). The lower cutoff frequency is due to the natural cutoff of the guide itself and the upper one is due to the succession of cavities in the filter section. Above the upper cutoff several pass bands occur at higher frequencies but these can be eliminated by cascading two or more filters of different characteristics.

D. TECHNICAL SESSION ON PROPAGATION

Paper #96. "Continuous Tropospheric Sounding by Radar", A.W. Friend, RCA.

Continuous film records of echoes received from vertically-aimed radar beams, were described. Heights up to 30,000 feet were recorded and a radar of about 1 Mw peak pulse power at 10 cm wavelength was used for some measurements.

Rapidly moving echoes from boundary discontinuities termed "angels" were observed to be most dense at the top of a moist layer of air. The equipment can "see" clouds not visible to the naked eye and records have often been made of cloud formations at great heights on an apparently clear day.

Dr. Friend stated that, as a general rule, the power required for thistype of investigation was about a megawatt and the most useful frequency range depended somewhat on the type of discontinuity. Roughly these frequencies were:

Precipitation echoes	-	up to 600 mc/sec
Dielectric Boundary changes	-	3000 - 5000 mc/sec
Fine Atmospheric dust	-	above 10,000 mc/sec

Paper #97. "A Theory on Radar Reflections from the Lower Atmosphere", W.E. Gordon, University of Texas.

Ordinary meteorological measurements seldom show sufficient change in air dielectric constant at air mass boundaries to explain the degree of radar reflections usually observed. New rapid response instruments have been developed which indicate surprising gradients of wind velocity, temperature, and humidity in regions of turbulence. A hot wire anemometer using fine platinum wire has been developed, and a differential resistance thermometer using two fine platinum wires stretched 6 mm apart has also been built. A rapid response humidity change recorder has been developed at the University of Chicago. Measurements with these instruments have shown changes which account fully for the dielectric constant changes expected from radar reflection.

Paper #98. "New Techniques in Quantitative Radar Analysis of Rainstorms", D. Atlas - Air Material Command, Wilmington, Ohio.

This paper described a rapid means of obtaining the approximate rainfall contours of a storm visible by radar.

Briefly, the scheme is to photograph the PPI picture of a rainstorm at several calibrated receiver gain settings. The photographs are made as a multiple exposure and on alternate receiver gain settings the PPI picture is made "positive" and "negative" in contrast. The resulting picture is a rough contour map of the storm with the values of rainfall intensity known if the receiver calibration is known. To partially correct for attenuation of energy being reflected from the far edge of the storm, the author suggests the use of two radar stations, one on each side of the storm. (This might work only on rare occasions, unless the co-operation of the storm could be enlisted.)

Paper #99. "Propagation of Waves Through the Ground, presented by K. McIlwain, Hazeltine Electronics Corp.

Experiments in propagation of radio waves of frequencies from .5 mc to 1000 mc/sec were carried out between two dry wells 200 feet deep. It was found that, due to abrupt changes in attenuation of a horizontal path at various levels, the earth behaved more as a set of lossy cables than as a homogeneous mass. Theoretical curves for attenuation versus depth in a homogeneous earth are expected to hold rarely in actual practice due to stratification.

E. TECHNICAL SESSION ON MICROWAVES

Paper #117. "Analysis and Performance of Waveguide Hybrid Rings", H.T. Budenborn, BTL.

This paper was an analysis of three and four arm rings. Equations were derived for the equivalent 2-mesh network of a three arm ring. To obtain practical results comparable with theory, tolerances on the hybrid ring had to be maintained to ±1/2 thousandth and 10 minutes of arc on the annulus. When this is done the attenuation on the conjugate paths is 48 db and the loss on zero db paths is .6 db. The usable frequency band of the hybrid ring is only ±5%. Results of this work will be published in the July, 1948 issue of BSTJ.

Paper #118. "Frequency Stabilization with Microwave Spectral Lines", presented by W.D. Hershberger, RCA.

Reducing the pressure of NH_3 gas to 10^{-2} mm of mercury in a microwave cavity can raise the effective Q to 100,000. The frequency of absorption can be determined to 1 part in 10^6, hence it is suggested that excellent stabilization could be obtained from such a gas cell. Fig. 10(a) shows a block diagram of a stabilized microwave system.

The difference frequency amplifier can be crystal-controlled if desired. Low frequency oscillators can be stabilized on harmonics. The presence of a foreign gas has the same effect on the Q as a higher pressure of NH_3 but has no effect on the frequency. Doppler effect places an upper limit on the Q obtainable by lowering the pressure. A maximum Q of 250,000 is predicted for ammonia but 500,000 could be obtained using the heavier gas, freon. Electric fields up to corona strength effect the frequency up to 17 mc/sec deviation in an NH_3 cell. Magnetic fields of 1000 gauss will move the line about 1700 kc. This work is to be published in an early issue of RCA Review.

Paper #119. "Analysis of a Microwave Absolute Attenuation Standard", by A.B. Giordano, Polytechnic Institute of Brooklyn.

Fig. 11(a) shows the type of attenuator discussed. The coaxial line is broken by a section of guide operating at below cutoff frequency. Transmission through this section is by induction fields only. A mathematical analysis of the system predicted linear attenuation with distance b, the length of the guide section. Measurements have shown this to be true with the exception of cases where b is very small. The attenuation curve is shown in fig. 11(b).

LECTURES ATTENDED BY MR. REDHEAD

"Physical Limitations on Directive Radiating Systems", L.J. Chu, M.I.T.

This paper discussed the relationship between radiation gain and impedance bandwidth of an arbitrary radiating system. An antenna with dimensions as shown in fig. 12(a) where (a) represents the radius of the enveloping sphere, may be represented by an equivalent network as shown in fig. 12(b).

The equivalent circuit of a dipole is shown in fig. 12(c), where (a) is the radius of the enveloping sphere, and (c) the velocity of light. The equivalent circuit of a TM_n spherical wave is shown in 12(d). The power in R is the power in the wave. The Q of the circuit (representing the impedance bandwidth) may be calculated by normal circuit techniques.

Different field distributions may be represented by taking different numbers of terms in the series for the equivalent circuit. Figs. 12(e) and (f) show qualitatively the effect of increasing a number of terms in the series (N). Any given antenna system may be represented by a series with an appropriate value of N.

"The Radiation Resistance of an Antenna in an Infinite Array of Waveguide", H.A. Wheeler, Consulting Radio Physicist, Great Neck, L.I. N.Y.

This paper is printed in full in the April issue of the Proceedings of the IRE.

"Reflectors for Wide-angle Scanning at Microwave Frequencies", R.C. Spencer, Wade Ellis and Ellen C. Fine, Watson Labs.

A purely theoretical discussion of scanning by means of a moving reflector inside a parabaloid was presented. Since this paper was largely mathematical it cannot easily be presented in a short report.

The optimum type of reflector was found to be a barrel with its axis perpendicular to the plane of scan and to the axis of the parabaloid.

"Mixed Impedance of Vertical Antennas over Fixible Ground Planes", A.S. Meier and W.P. Summers, Ohio State University, Columbus, Ohio.

This paper reported measurements of the variation of radiation resistance of a unipole above a small metallic ground plane where the size and the shape of the plane was varied. The dimensions of the ground plane were relatively small in relation to wavelength.

The input impedance of the unipole was found to be a damped oscaillating function of wavelength and ground-plane dimensions. The impedance with a circular ground plane was found to vary from ±5% to ±20%. With a square ground plane the variations were about 50% of those with a circular ground plane.

In general, the antenna thickness was found to have but small effect on the impedance but size and shape of the ground plane had a profound effect.

"Current Distributions on Aircraft Structures",
J.V.N. Granger, Harvard University, Cambridge, Mass.

The experimental technique for measurement of R.F. current distribution on aircraft structures was presented.

A model, sealed-down twenty times was used in these experiments and a wavelength of 10 cm. A small hand-operated search coil with a built-in crystal detector was used feeding into a d-c amplifier.

Measurements on a P-47 aircraft with a v.h.f. stub revealed that the currents flowing in the wings are in antiphase and so contribute nothing to the resulting field distribution. Some attempt was made at correlation between the number of lobes in the field pattern and the current distribution on the surface of the aircraft. This correlation was not at all convincing.

"Thermionic Emission from Grids in Vacuum Tubes",
M. Arditi and V.J. DiSantis, Federal Telecommunication Labs. Inc., Nutley, N.J.

Emission from control grids of vacuum tubes is a serious limitation in tube design. This paper considered thermionic emission only.

Barium oxide evaporates from the cathode into the grid, the material of the grid "alloys" with the BaO to form a substance of high work function.

With the control grid negative dissociation of the grid surface can only occur when the grid temperature is above the dissociation temperature of the material. With positive grid voltages, dissociation occurs due to the kinetic energy of the bombarding electrons even though the temperature of the grid is well below the thermal dissociation temperature.

The critical voltages at which dissociation occurs may be found from the heat of formation of the substance. Molybdenum oxide is 7.5 Ev and platinum oxide 0.74 Ev. The probability of dissociation by fast electrons is less than that for slow electrons, it is also dependent on the rate of diffusion of oxygen through the oxide coating.

Gold plating was found to reduce grid emission but the vapourized gold was found to cause poisoning of the cathode.

Molybdenum oxide was found to cause poisoning at 7.5 Ev as predicted.

Tungsten carbide did not cause poisoning and the grid current was very low.

Boron carbide was satisfactory on a molybdenum grid, the boron carbide has a contact potential of 1.1 to 1.5 v.

It was also found that grid emission rises with time when a thoriated tungsten filament is used.

The most satisfactory material for coating grids appears to be boron or tungsten carbide.

"The Negative Ion Blemish in a Cathode Ray Tube and its Elimination", R.M. Bowie, Sylvania Electric Products, Inc., Flushing, N.Y.

The ion blemish in C.R.T.'s is a cause of great annoyance in television tubes. All the ions appear to start from the cathode, three methods of elimination are used -

(1) Proper processing of tube to remove all traces of gas.
(2) Filter laid against screen consisting of thin film of metal, collodion, etc. Penetration is proportional to $\frac{E^2}{m}$ so that only a few ions can penetrate.
(3) Use of a negative ion trap which depends on the fact that the ions are relatively unaffected by magnetic fields, and may be separated from the electron beam by some type of trap similar to those shown in figs. 13(a,b).

"Wide-Tuning-Range Continuous-Wave High-Power Magnetrons", P.W. Crapuchettes, Letton Industries, San Carlos, California.

The construction of a 1 kw magentron capable of ±20% tuning at 10 cm was described. The efficiency was reasonably flat over the range 8 - 13 cm at about 63%. Fig. 14 shows the way in which the tube changed from one mode to another.

"Wide-Range Tuning Systems for Magnetrons", E.N. Kather, Raytheon Manufacturing Co., Waltham, Mass.

The various methods of tuning magnetrons were described.

(1) Coupled external reactance gives 10% tuning without moding.
(2) Capacity tuning ("cookie-cutter").
(3) Vane type tuner with fingers extending into the capacitive portion of the cavities.

(4) Inductive tuning with fingers into inductive portion of cavities. This system has greater losses than the others.
(5) Capacitive and inductive tuning, tuning range 1.9 to 1.

Allis has shown that $V \propto \frac{B}{\lambda}$. A constant operating voltage can be obtained by using the tuner to vary the flux. Also the efficiency parameter $\mathcal{D} \propto \sqrt{B\lambda}$, by suitable adjustment of the output coupling to give $\eta_c \propto \frac{1}{\lambda}$ (the current efficiency) a constant efficiency may be obtained.

A resonance of the tuning fingers was observed which was similar to the magnetron resonance. A loss in power output and efficiency occurs at points of intersection of the tuning curves of the resonators proper and the tuning fingers.

"New Design for a Secondary Emission Trigger Tube NU-TR-1032-J",
C.F. Miller and W. McLean, National Union Radio Corporation, Orange, N.J.

Fig. 15(b) shows the V-I characteristic of the dynode system. The tube will stabilize at Y when B^+ disconnected.

The tube construction is shown in 15(c). A_3 shields A_1. The dynode (D) was made of OFHC copper with a coating of BaO and MgO. A_3 can be used as a modulator electrode. Fig. 15(d) shows the tube in use in a square wave generating circuit.

0.2 μs rise-time was obtained on square wave using this circuit.

"A Spiral-Beam Method for the Amplitude-Modulation of Magnetrons",
J.S. Donat and R.R. Bash, RCA, Princeton, N.J.

A beam of electrons spiralling in a longitudinal magnetic field varies the conductance presented by a resonant cavity coupled to a magnetron and thus varies the power delivered to the load. (See fig. 16). A pure conductance is presented by the absorption tube at the cyclotron frequency for a beam parallel to the magnetic field. A voltage modulation of 85% is obtainable with fair linearity and a bandwidth of at least 20 mc/s while the frequency modulation is only ±15 kc.

50 - 100 ma is drawn in the absorption tube. 0.4 watts are required to modulate 1/2 kw. The power limit is the collapse of the glass walls of the tube due to absorption of R.F.

"The Dyotron - A New Microwave Oscillator",
E.D. McArthur, General Electric, Schenectady, N.Y.

This tube is a modification of the disc seal triode 2C39. A silvered mica condenser is inserted between grid and cathode assemblies in the tube. It is used in a co-axial resonator.

The expression for the anode current in a triode is given by:

$$i = K \left(\frac{E_b - e_p \sin wt}{\mu} + E_G + eg \sin wt \right)^{3/2}$$

the last term is automatically zero in the dyotron and thus,

$$i = I_c - \frac{A_{ep}}{\mu} \sin (wt - N\pi)$$

where $N\pi$ represents the transit-time cathode to grid
Conditions for oscillation are

$$Q > 750 \, N_\mu \, \frac{Sg}{Sp}$$

N = 1,3,5 have been used so far in tuning ranges 370 to 3,700mc and 18 - 2,800 mc/s with V_a = 300 v I_a = 80 mA and an output of 300 - 350 mw.

"Electrostatically Focused Radial-Beam Tube",
A.M. Skellett, National Union Radio Corp., Orange, N.J.

This tube is an electrostatically focused switch tube similar in all respects, except focusing, to the magnetically focused tube previously reported in the PIRE. The radial beam is focused and rotated by six circumferential anodes connected in pairs to a three-phase supply.

"A New Two-Terminal High-Voltage Rectifier Tube",
G.W. Baker, Chatam Electronics Corp., New, N.J.

This rectifier tube is especially designed for high voltage R.F. power supplies where the heaters of the diodes are liable to be at very high d-c potentials with respect to ground. Thus the problem of heater transformers is extremely difficult. These tubes have a very thick cathode coating on a very fine base material and are heated by R.F. losses in the oxide coating. Thus no heater transformers are required.

"An Electronic Instrument for the Determination of the Dead Time and Recovery Characteristics of Geiger Counters",
L. Costrell, National Bureau of Standards, Washington, D.C.

Fig. 17(a) shows the operating of a Geiger counter for pulses close together in tune. If N is the number of counts and a the counting rate,

then, $N = a(1 - E^{at})$

if t_g is time that amplifier is gated and t_d is the dead time, then,

$$N_{(td)} = a\ (1 - Ea^{td})$$

Therefore, $N = a(E^{atg} - E^{atd})$

From the above it can be shown that

$$N = a^2\ (tg - td)$$

$$td = tg\ \frac{N}{a^2}$$

Fig. 17(b) shows the circuit used.
Fig. 17(c) shows the effect of varying the argon pressure in a G.M. counter.

"Swept Frequency 3-Centimeter Impedance Indicator",
H.J. Riblet, Submarine Signal Co., Boston, Mass.

Fig. 18 shows the circuit of this device. The "wave-sampler" consists of a wave guide with two slots in opposite walls at right-angles feeding into two separate guide systems.

"The Automatic V.H.F. Standing Wave Ratio Plotting Device",
W.A. Faiss, L.L. Mason and K.S. Packard.

This device measured S.W.R. by means of a directional coupler feeding an amplifier which measures the ratio of forward and reflected power by conversion of amplitude at the detectors to phase difference which operates a trigger circuit used as a d-c amplifier.

"Microwave Impedance Bridge",
M. Chodorow, E.L. Ginzton and J.R. Kane, Stanford University, Calif.

This paper described a 12-terminal wheatstone bridge network for use at microwave frequencies.

The balance conditions for a wheatstone bridge with arms A,B,C and D, is

$$A = \frac{B}{C} D$$

If D consists of a variable reactance and a matched resistance load, and B and C pure reactances so that the ratio B/C is a pure number, then by varying the reactive component of D and the ratio B/C any value of resistive and reactive components can be matched.

The V.S.W.R. measured by this system is as accurate as with the best slotted lines available. It has great advantages over the magic T network since any value of resistive load may be measured.

Fig. 19 shows the arrangement of the guide system forming the bridge.

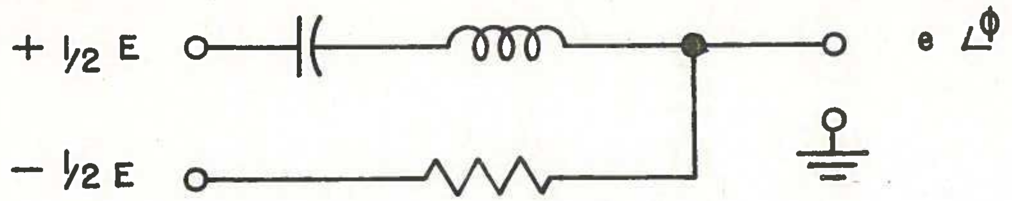
"<u>A Waveguide Bridge for Measuring Gain at 4,000 Mc,</u>"
A.L. Samuel, University of Illinois, Urbana, Ill.
C.F. Crandell, Southwestern Bell Telephone Co.

This paper described a bridge for the measurement of 4-terminal transfer impedance in the vicinity of 4,000 mc. This system has no novel techniques.

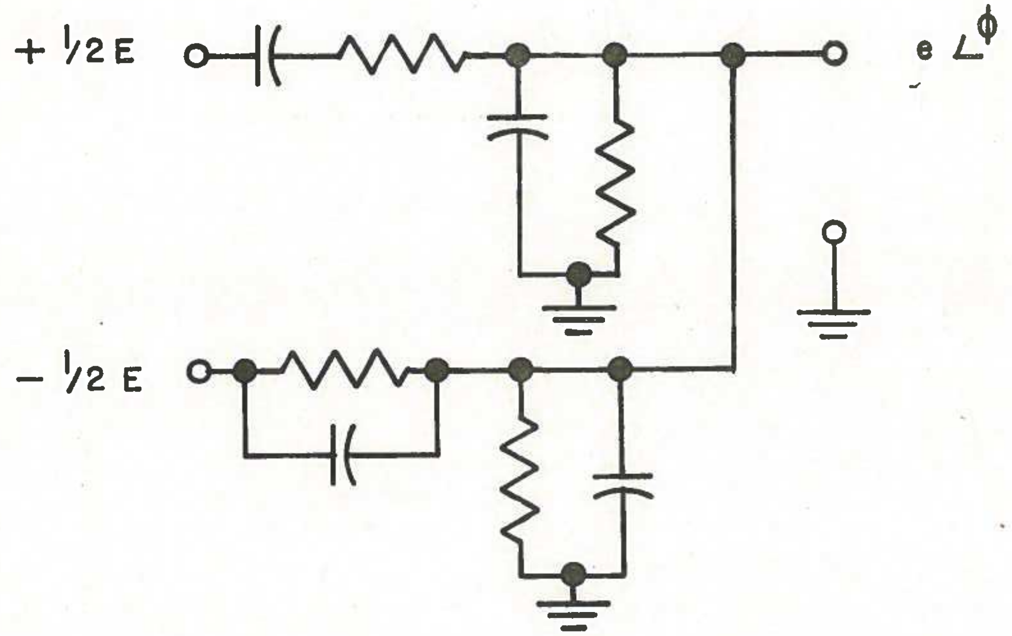
A block diagram of the system is shown in Fig. 20.

FIGURES

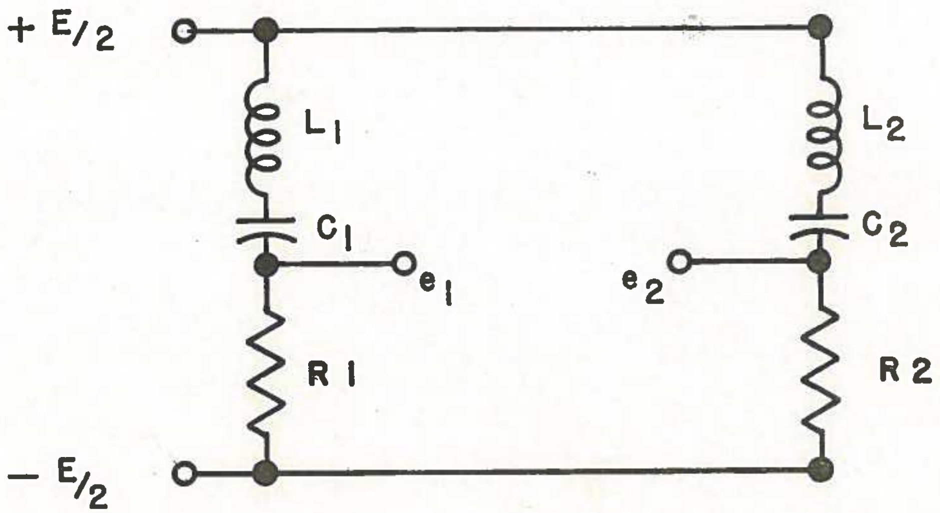
1(a)



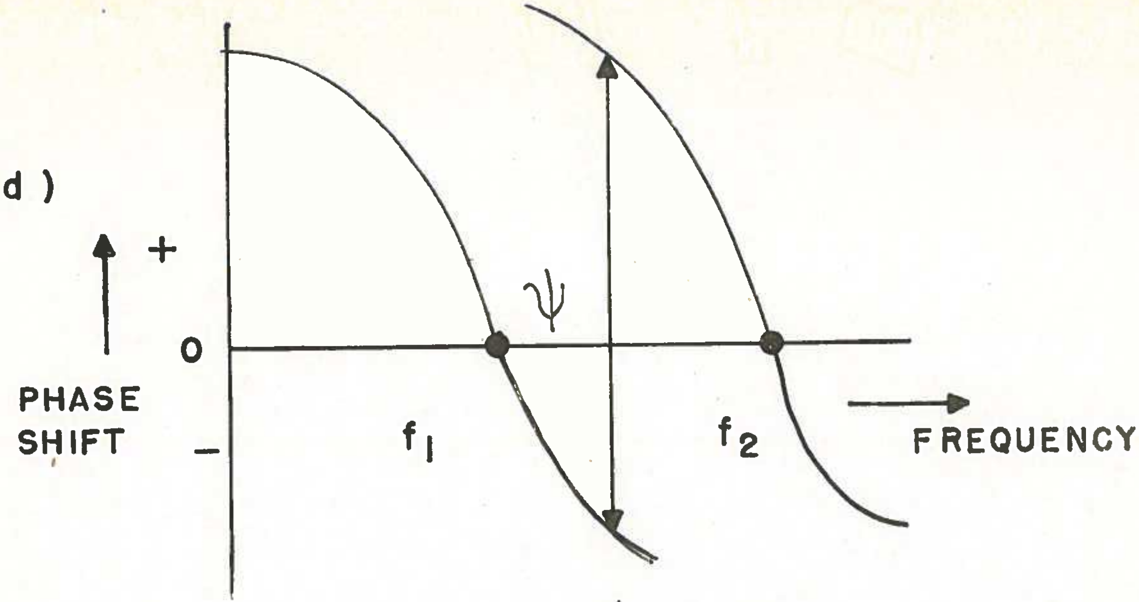
1(b)



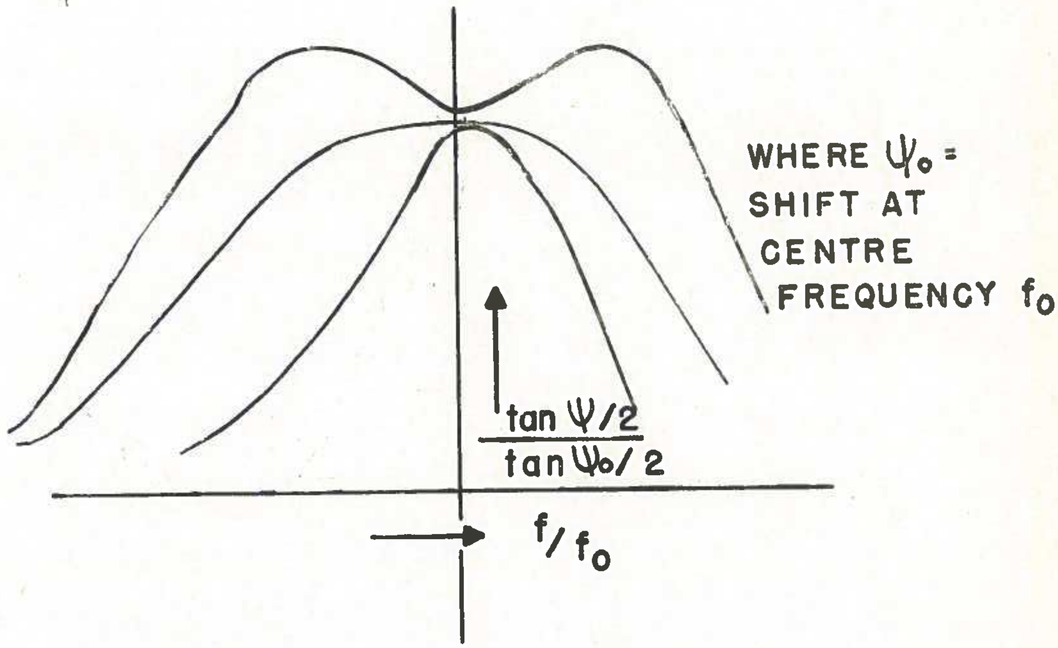
1(c)



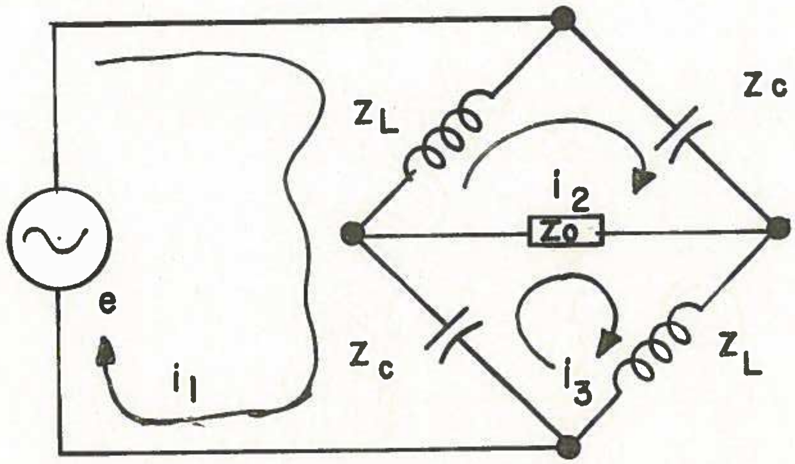
1(d)


+
0
PHASE SHIFT
−
f_1
ψ
f_2
FREQUENCY

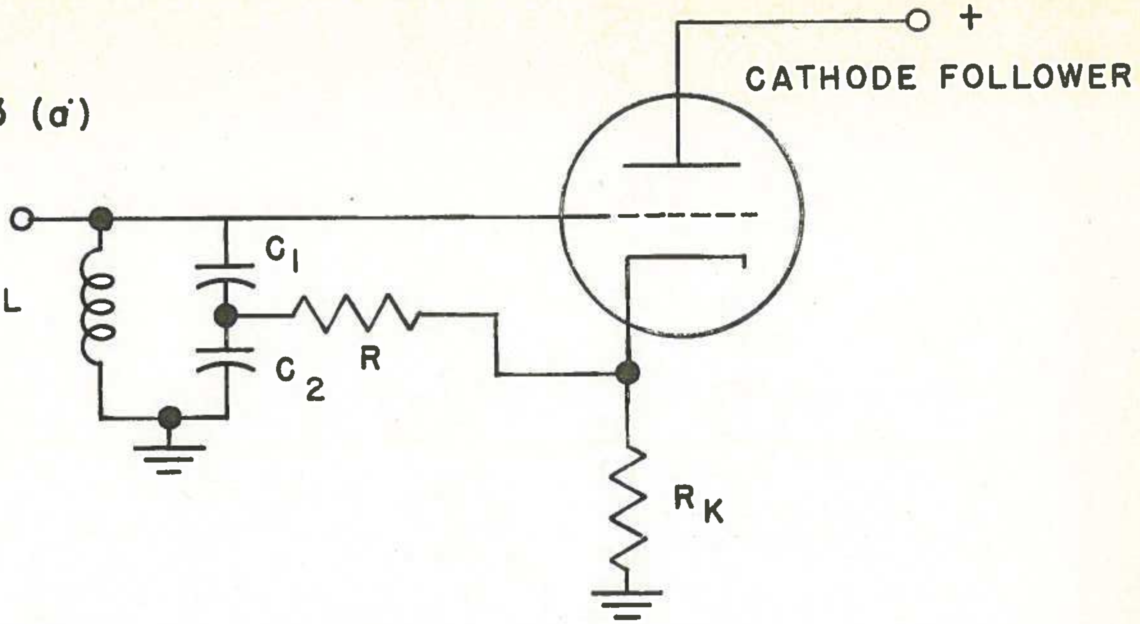
1 (e)


WHERE ψ_0 = SHIFT AT CENTRE FREQUENCY f_0
$\frac{\tan \psi/2}{\tan \psi_0/2}$
f/f_0

2 (a)

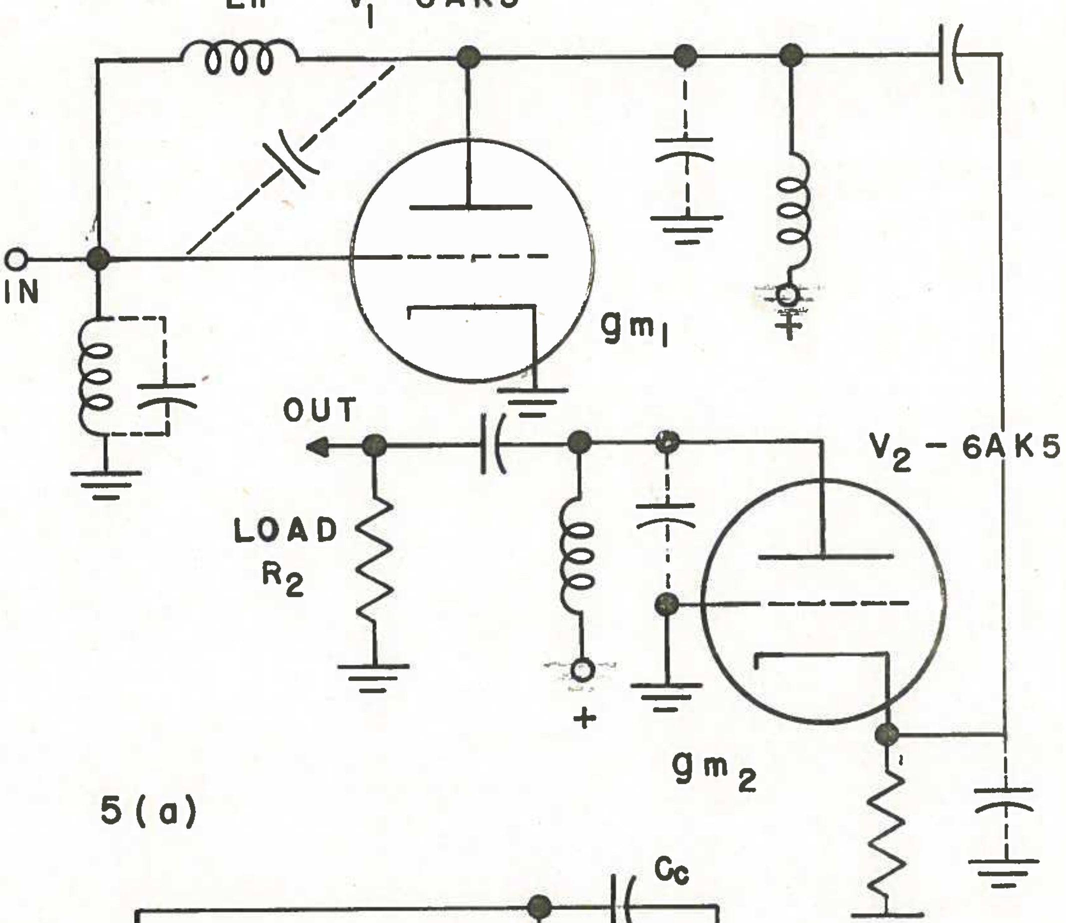

Z_L
Z_c
i_2
Zo
e
i_1
Z_c
i_3
Z_L

3 (a')



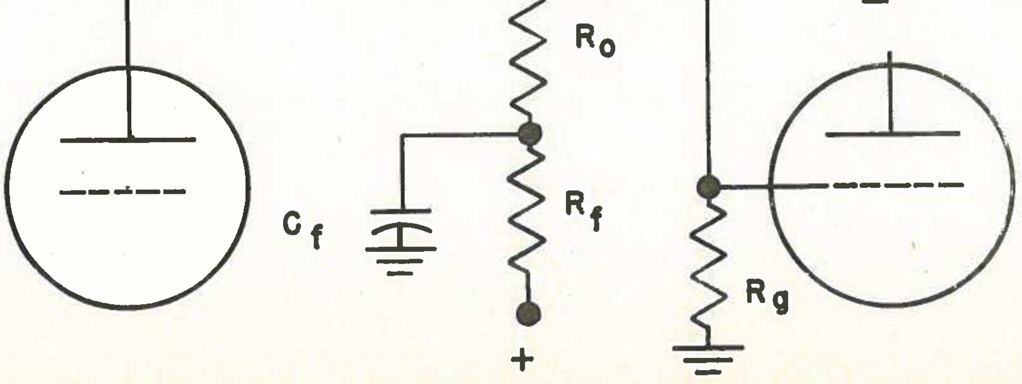
+
CATHODE FOLLOWER
L
C_1
C_2
R
R_K

4 (a)



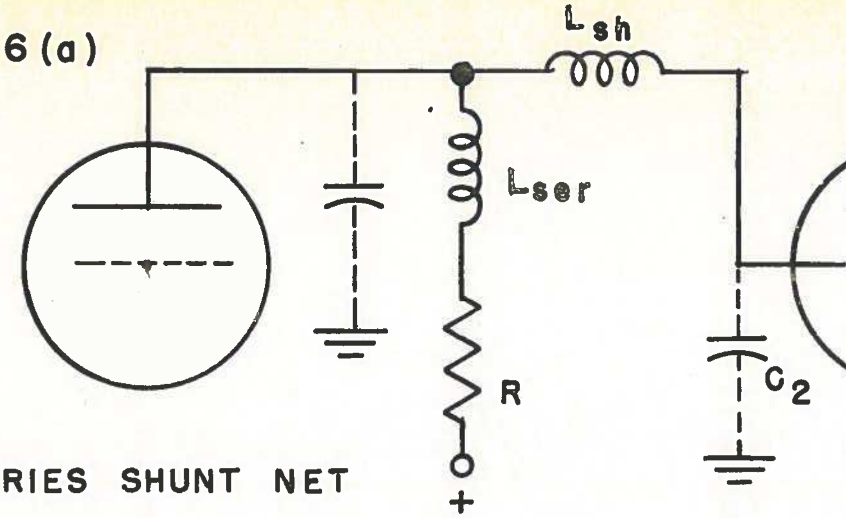
Ln
V_1 -6AK5
IN
gm_1
+
OUT
LOAD
R_2
+
V_2 - 6AK5
gm_2

5 (a)



Cc
R_o
C_f
R_f
R_g
+

6 (a)



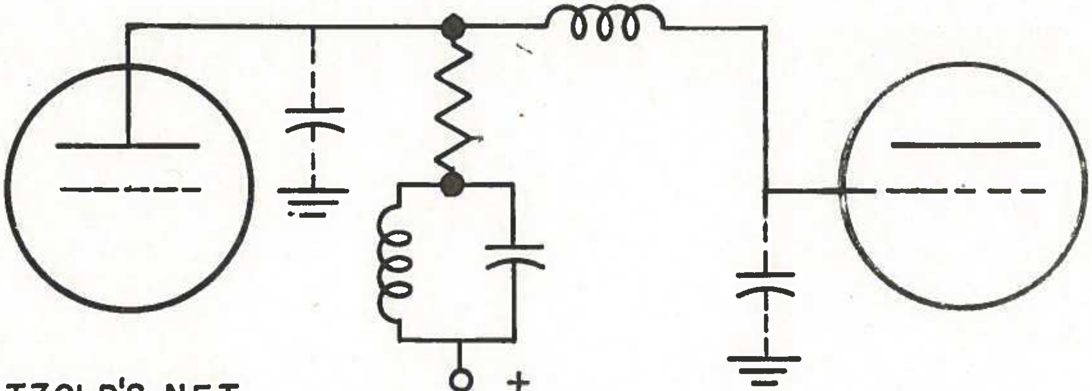
SERIES SHUNT NET

$L_{sh} = .7\, C_2 R^2$

$L_{ser} = .2\, C_2 R^2$

$F = 700$

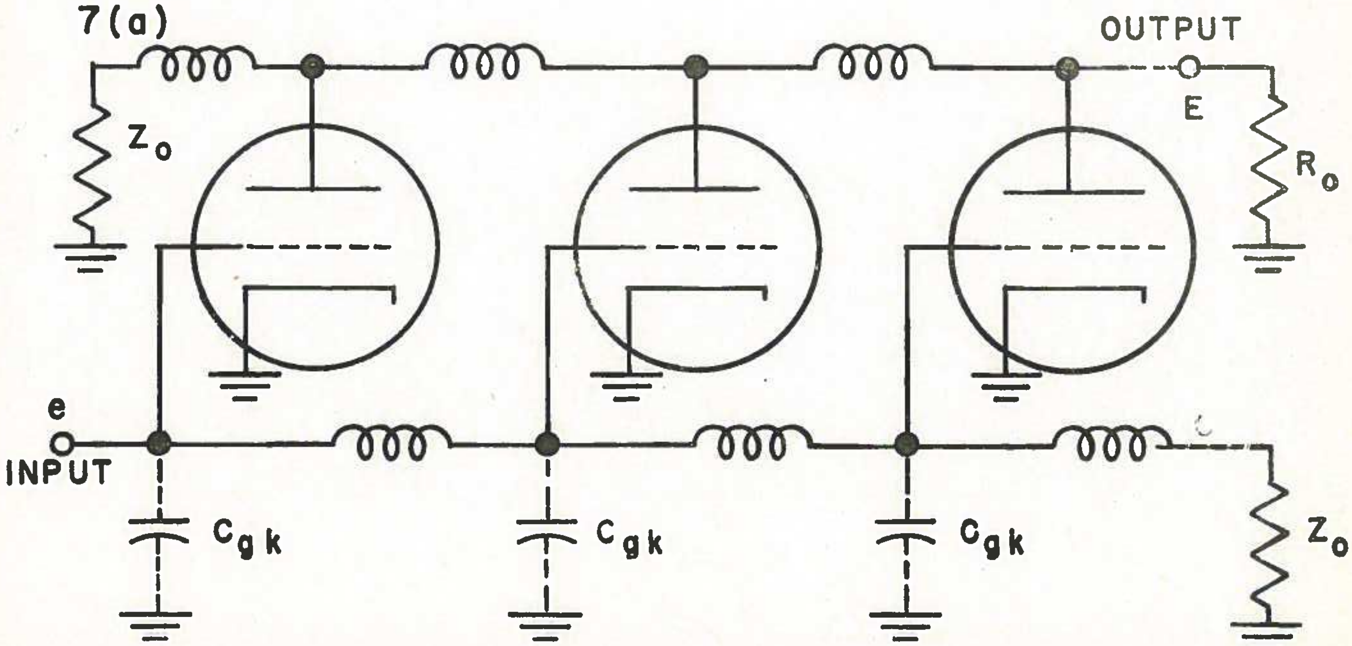
6 (b)



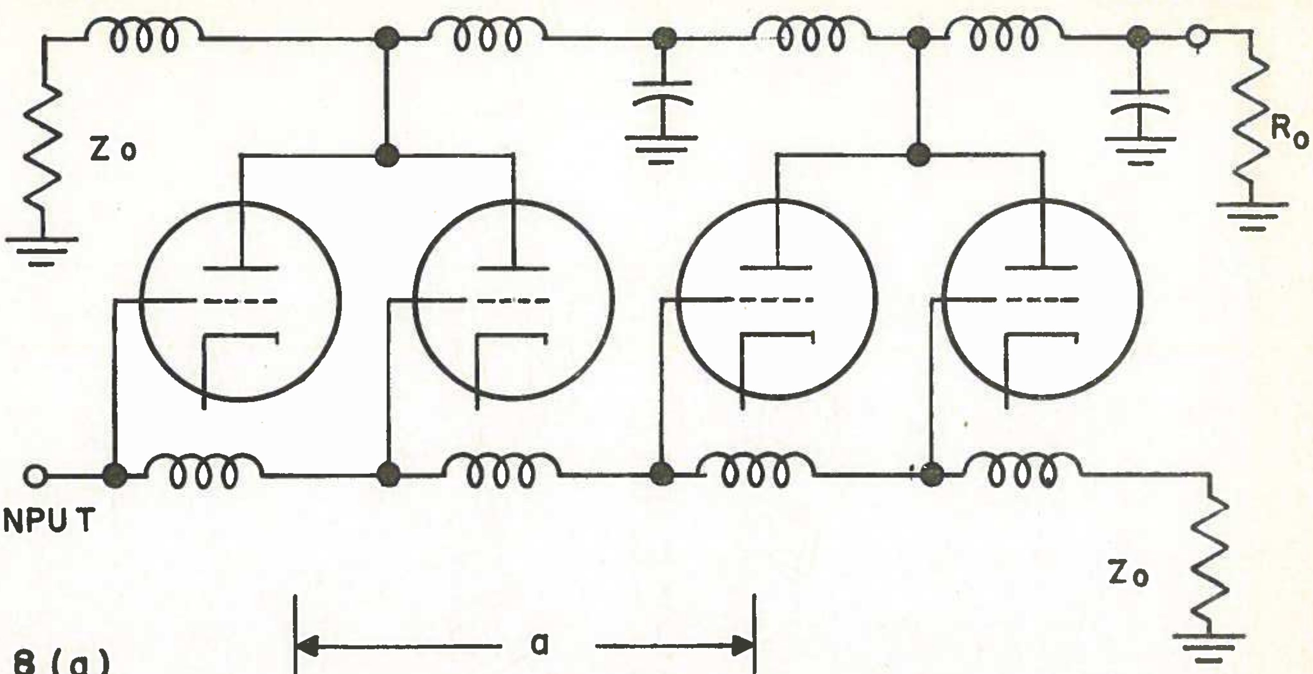
DIETZOLD'S NET

F = 1128

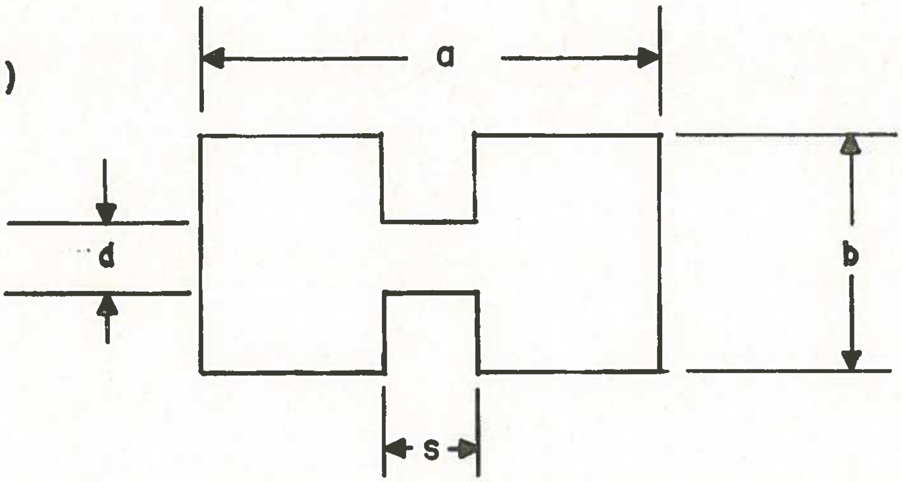
7 (a)



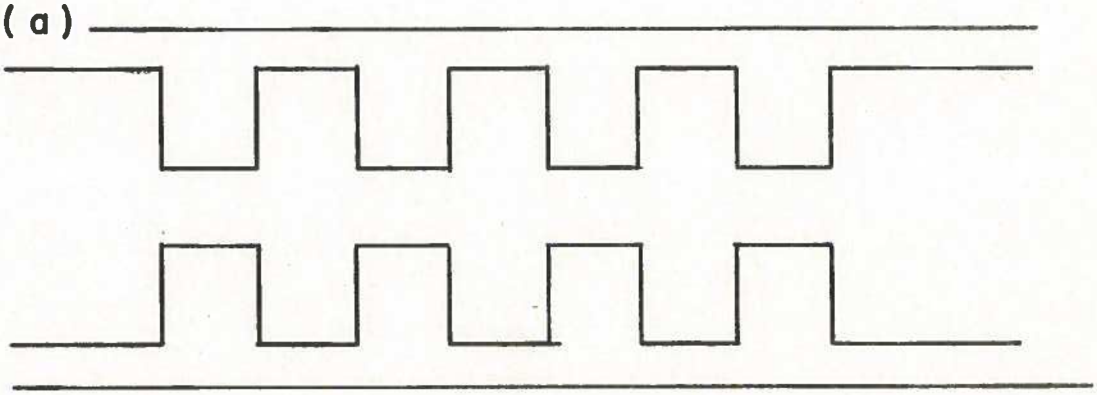
7 (b)



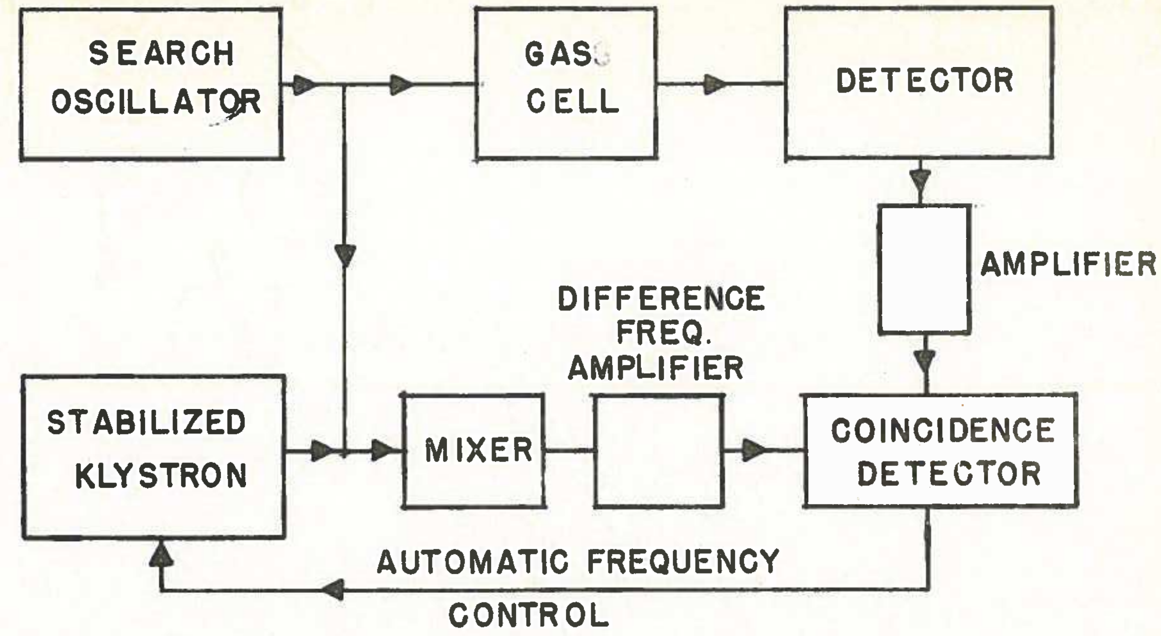
8 (a)



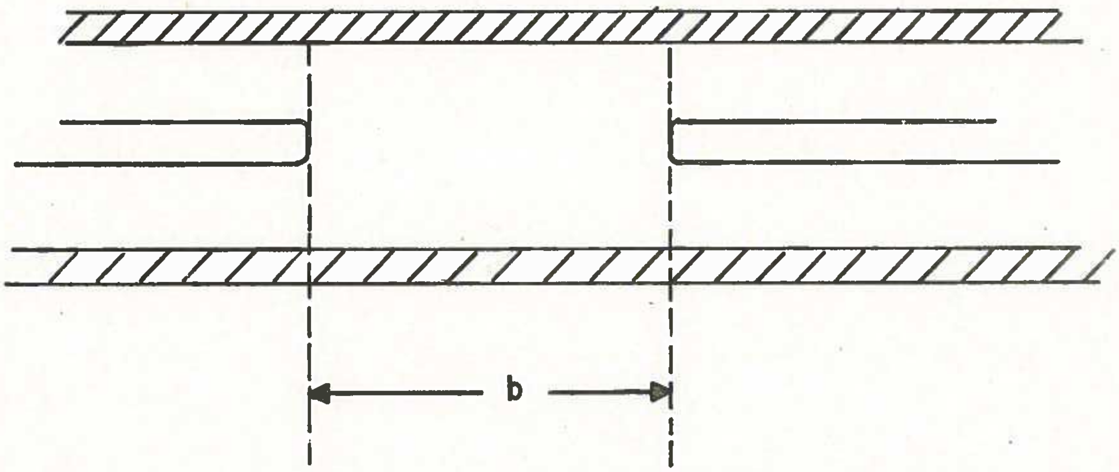
9 (a)



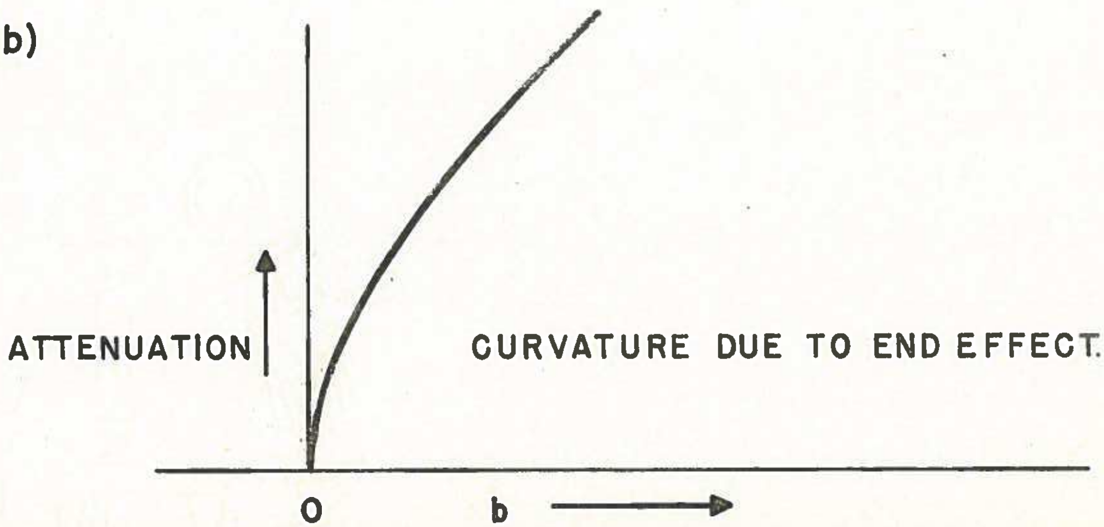
10 (a)


SEARCH OSCILLATOR
GAS CELL
DETECTOR
AMPLIFIER
DIFFERENCE FREQ. AMPLIFIER
STABILIZED KLYSTRON
MIXER
COINCIDENCE DETECTOR
AUTOMATIC FREQUENCY CONTROL

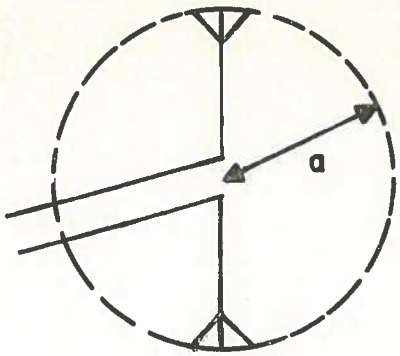
11 (a)


b

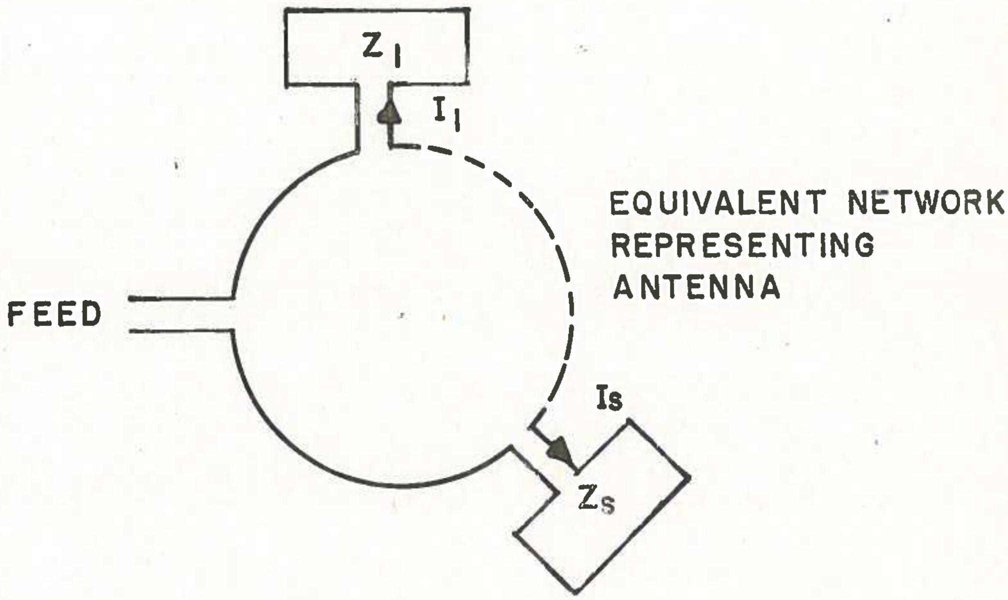
11 (b)


ATTENUATION
CURVATURE DUE TO END EFFECT.
O
b

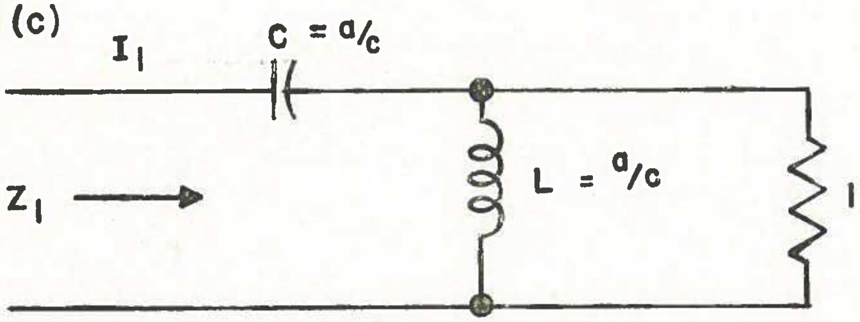
12 (a)



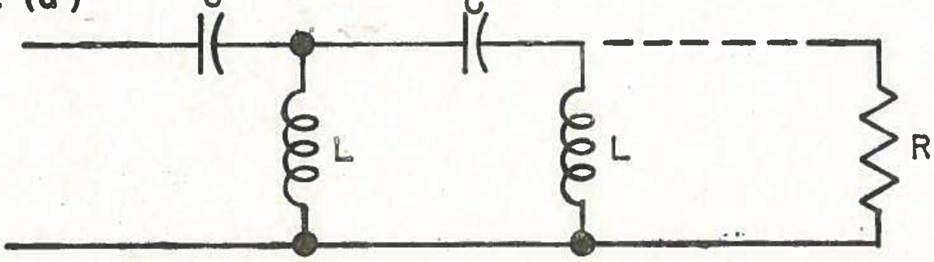
12 (b)



12 (c)

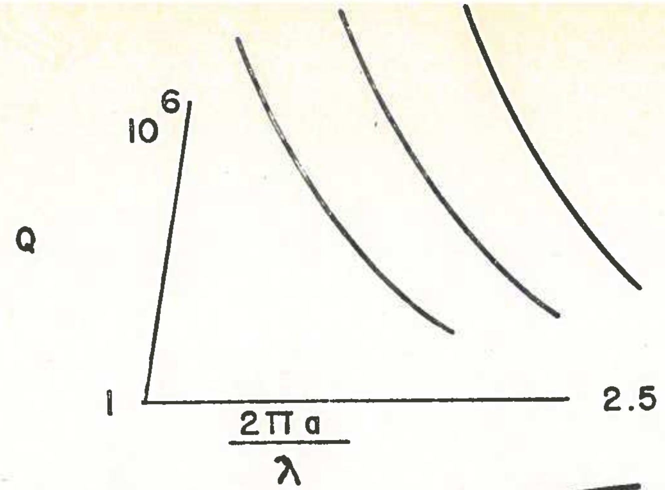


12 (d)

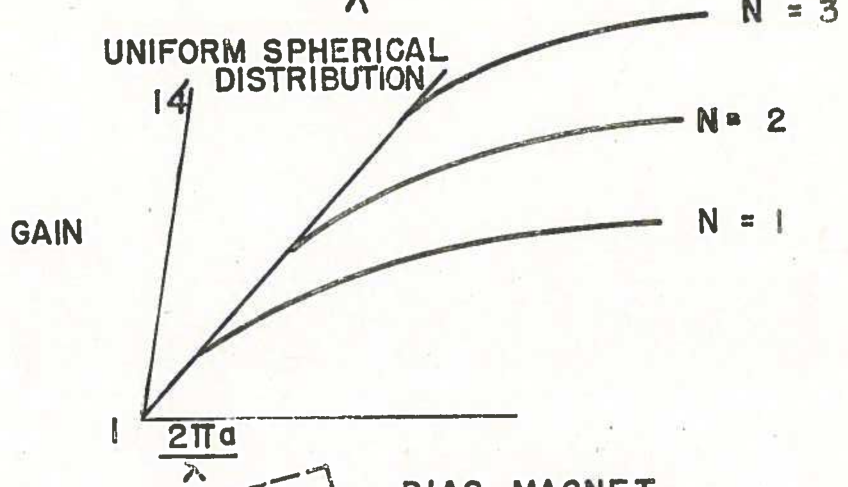


$$C = \frac{a}{nc}\,\frac{9}{(2n-3)e} \qquad L = \frac{9}{(2n-3)e}$$

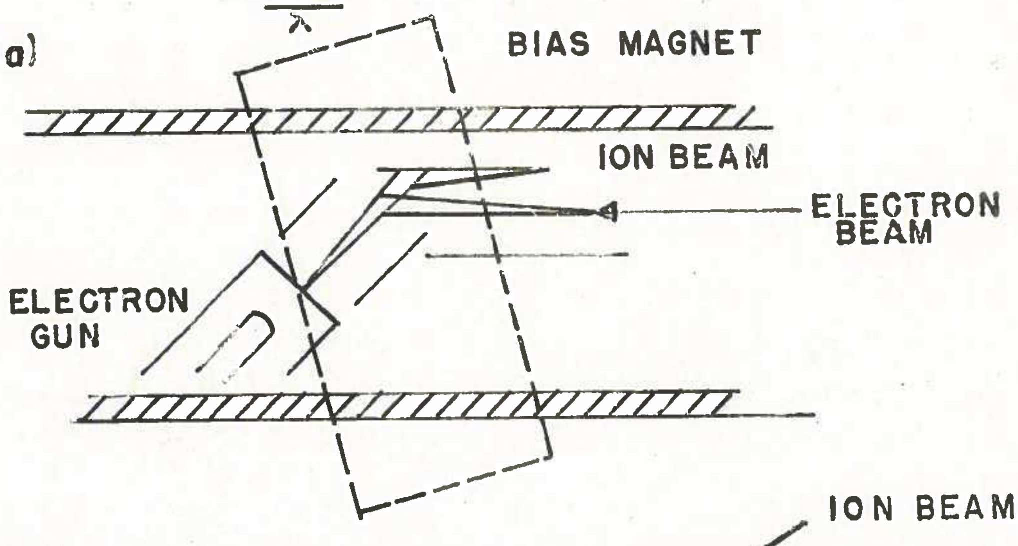
12 (e)



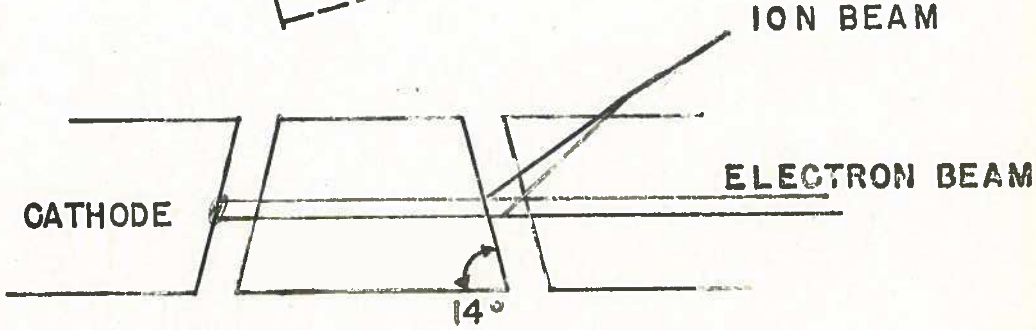
12 (f)



13 (a)

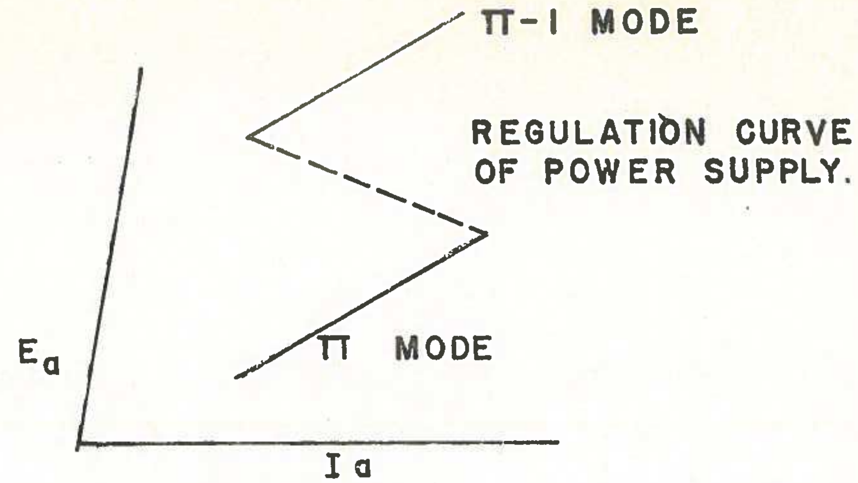


13 (b)

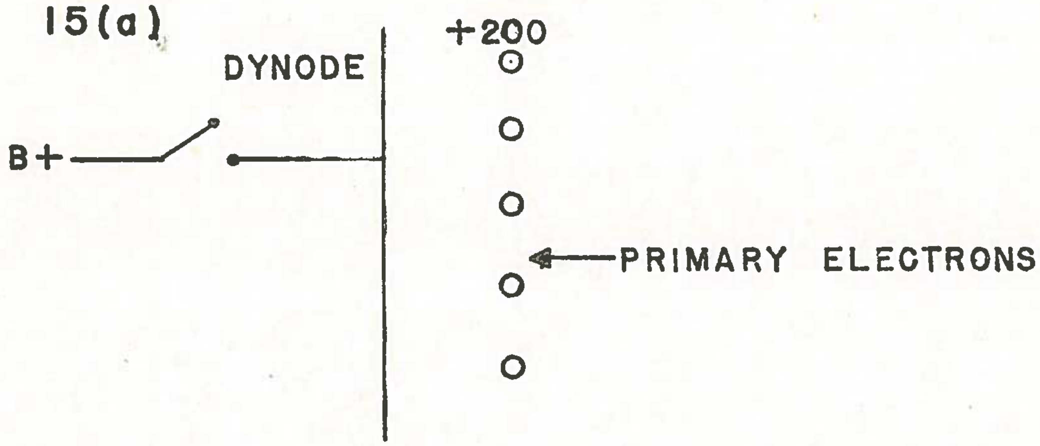


MAGNETIC FIELD COMPENSATES EFFECTS OF ELECTRIC FIELD ON ELECTRONS BUT IS NOT SUFFICIENT TO EFFECT IONS.

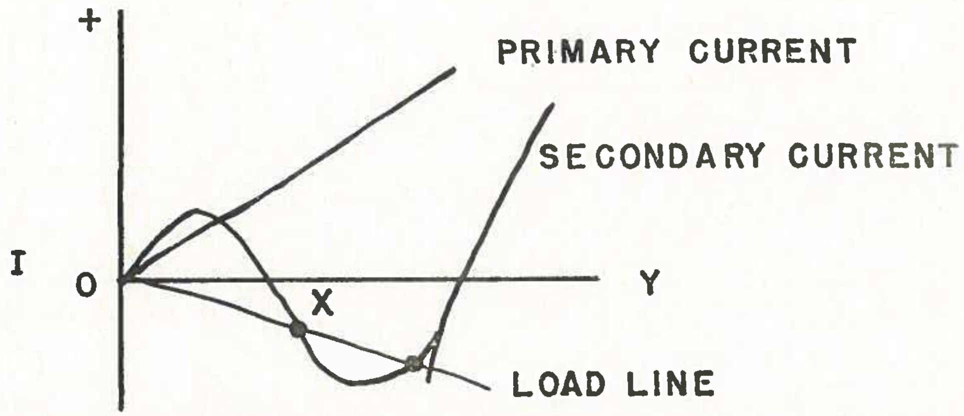
14



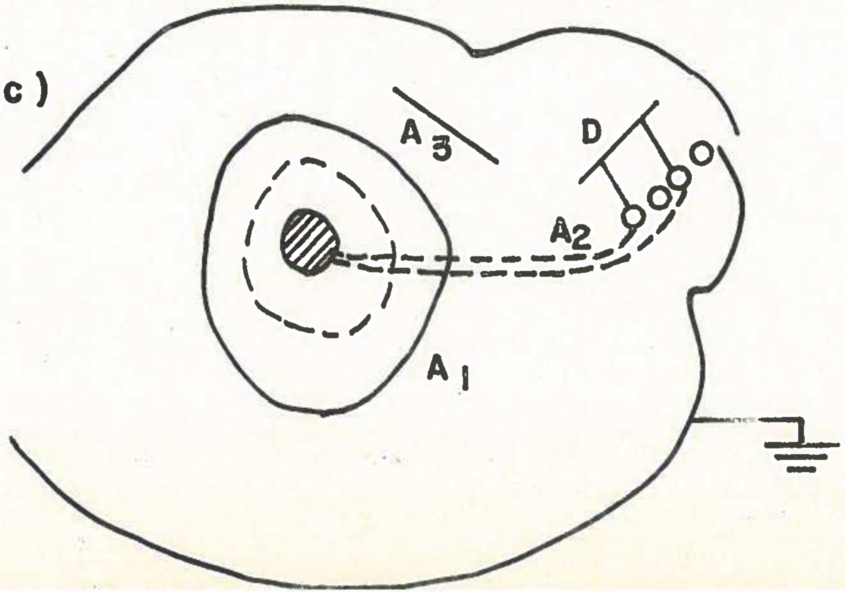
15(a)



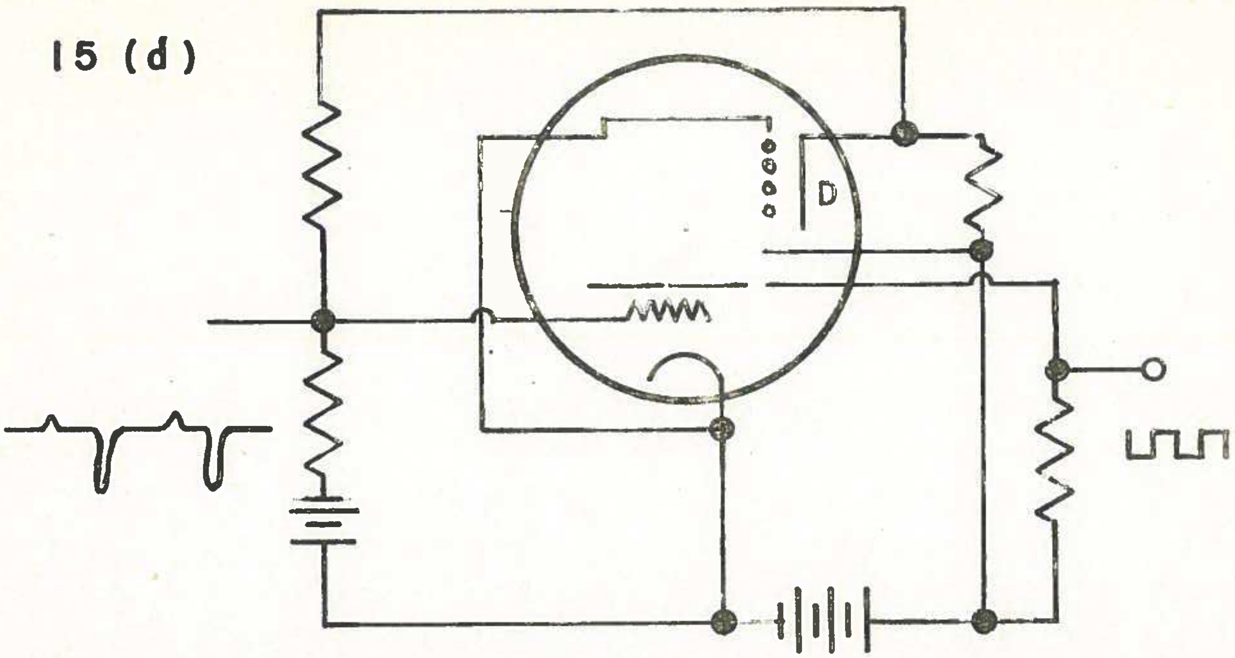
15(b)



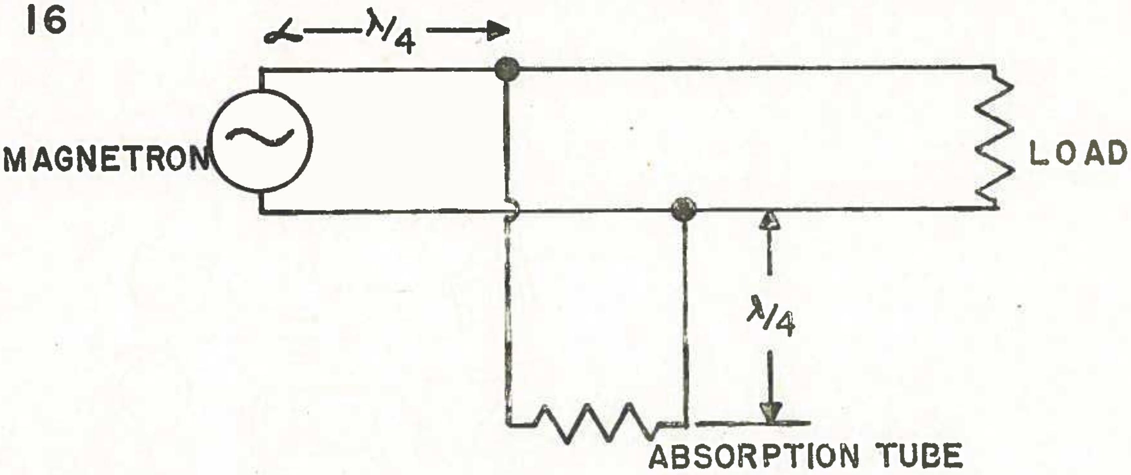
15(c)



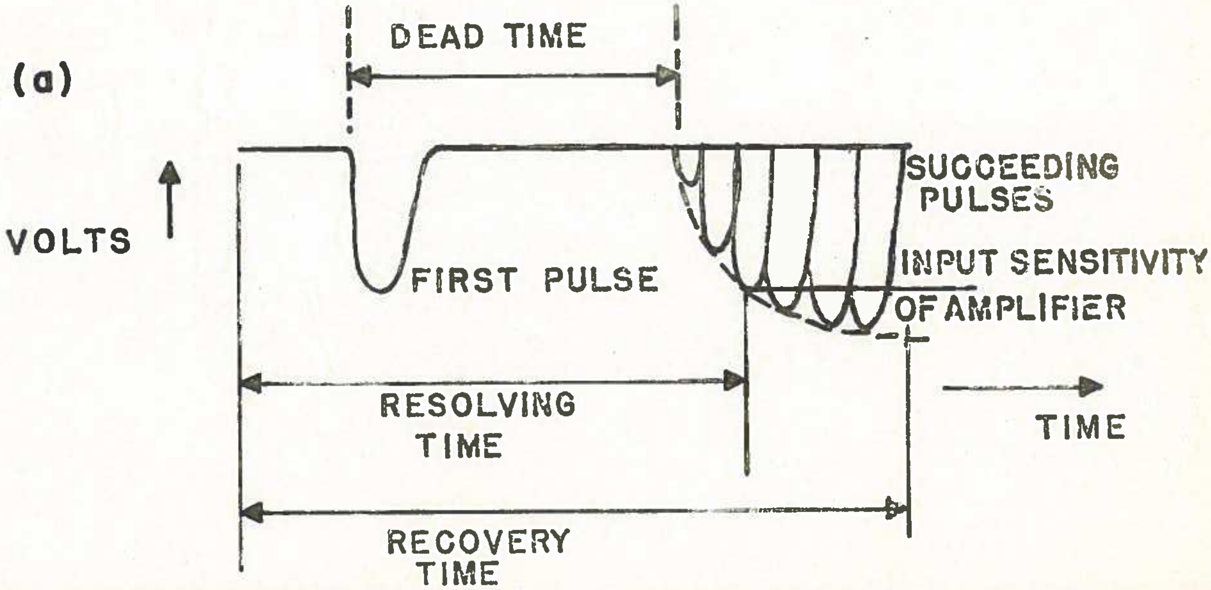
15 (d)


D

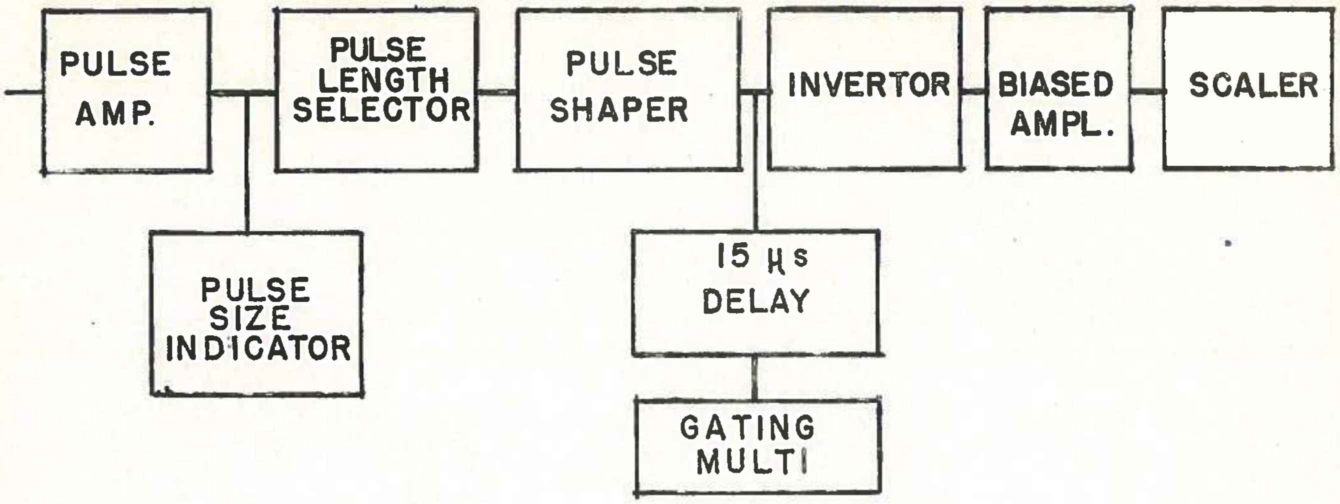
16


λ/4
MAGNETRON
LOAD
λ/4
ABSORPTION TUBE

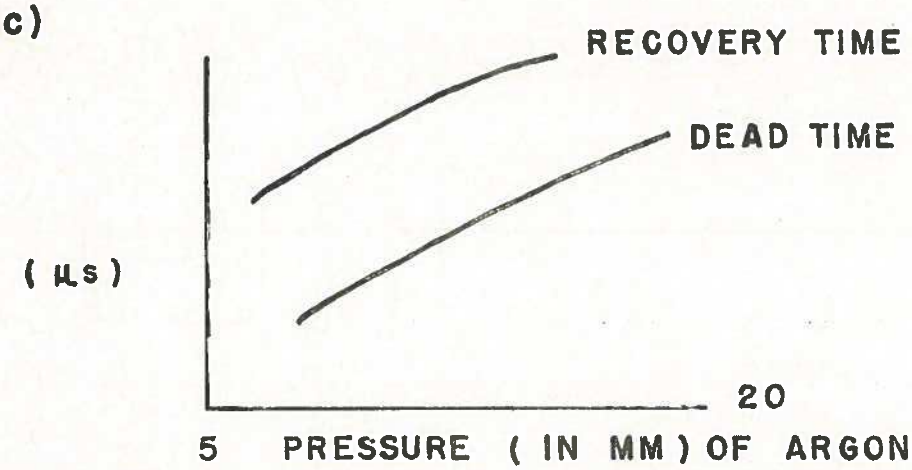
17 (a)


DEAD TIME
VOLTS
FIRST PULSE
SUCCEEDING PULSES
INPUT SENSITIVITY OF AMPLIFIER
RESOLVING TIME
TIME
RECOVERY TIME

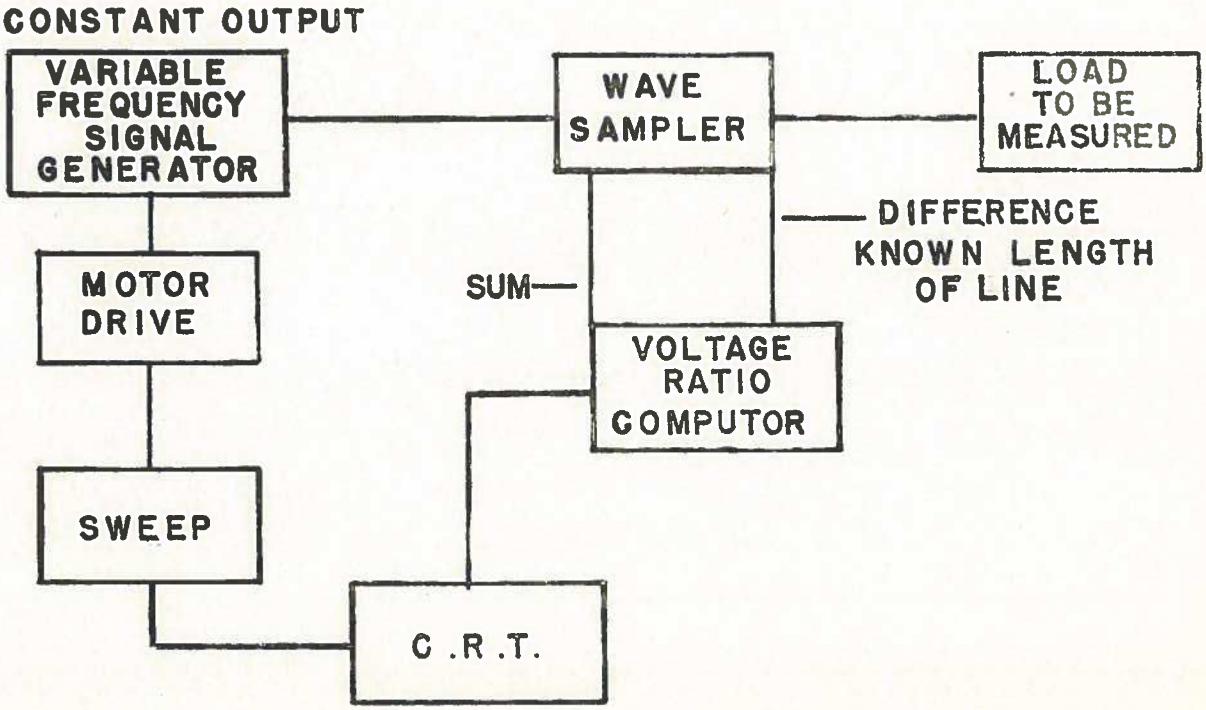
17 (b)


PULSE AMP.
PULSE LENGTH SELECTOR
PULSE SHAPER
INVERTOR
BIASED AMPL.
SCALER
PULSE SIZE INDICATOR
15 μs DELAY
GATING MULTI

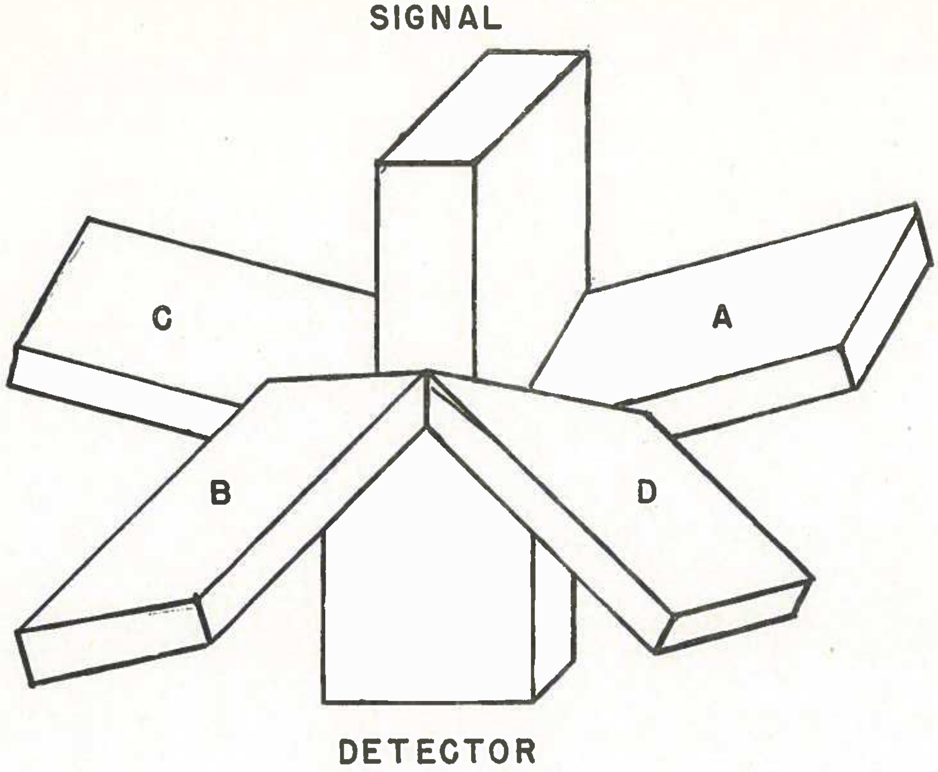
17 (c)


RECOVERY TIME
DEAD TIME
(μs)
20
5 PRESSURE (IN MM) OF ARGON

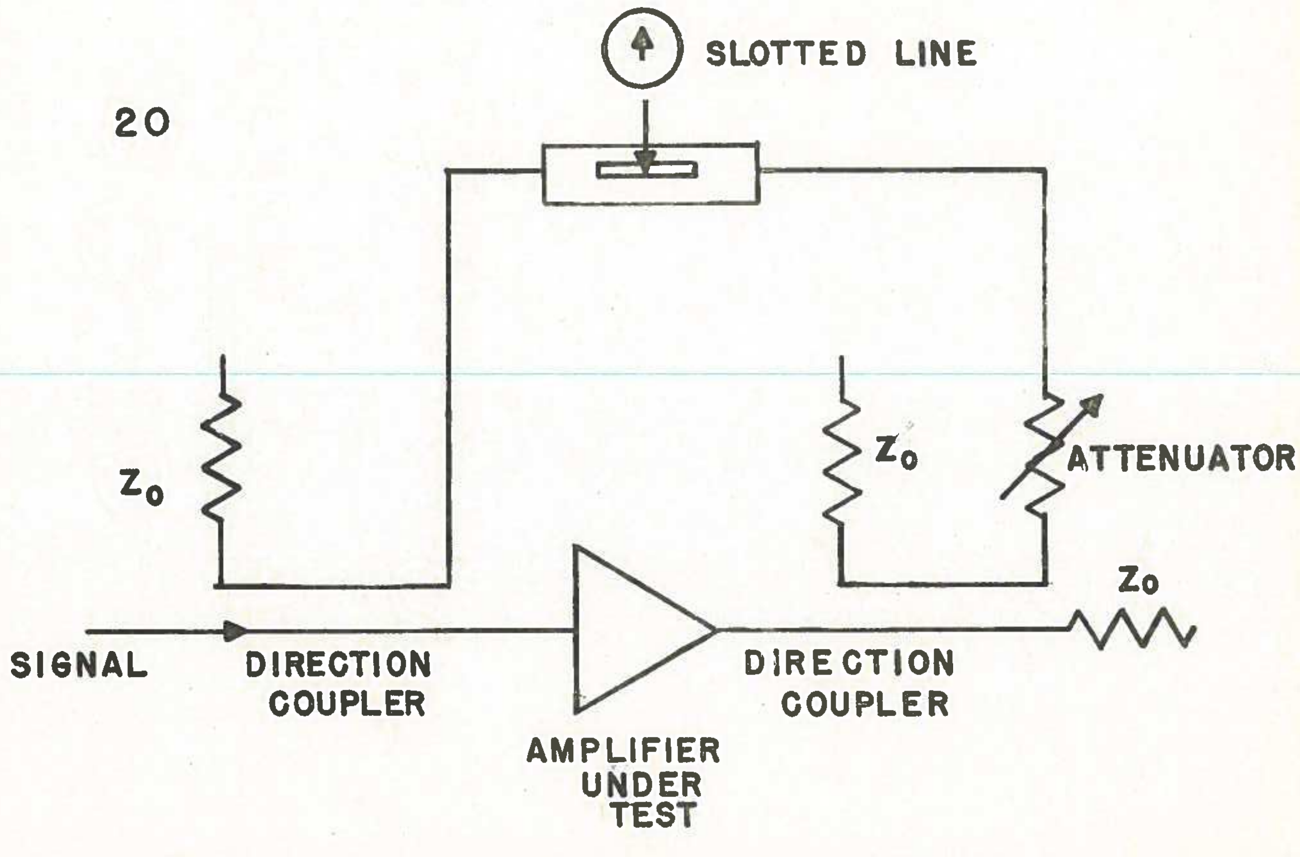
18.


CONSTANT OUTPUT
VARIABLE FREQUENCY SIGNAL GENERATOR
WAVE SAMPLER
LOAD TO BE MEASURED
DIFFERENCE
KNOWN LENGTH OF LINE
SUM
MOTOR DRIVE
VOLTAGE RATIO COMPUTOR
SWEEP
C .R .T.

19


SIGNAL
C
A
B
D
DETECTOR

20


SLOTTED LINE
Z_0
Z_0
ATTENUATOR
Z_0
SIGNAL
DIRECTION COUPLER
AMPLIFIER UNDER TEST
DIRECTION COUPLER