in terms of density but in terms of the number and nature of the tracks recorded by nuclear particles, namely, electrons, mesons, protons, deuterons, alpha particles, fission fragments. The tracks are for the most part a linear array of silver grains, and a fine grained emulsion of uniform sensitivity—grain to grain—is required. In the recording of high-energy protons, for example when the ionization per grain hit is low, the highest sensitivity and maximum development is required. On the other hand, for the selective recording of heavy ionizing fission fragments a plate of low sensitivity is required in order that the other radiations of lesser ionization may not increase the background of the plate. In general the tracks of nuclear particles in the photographic plate are evaluated in terms of their length, grain spacing, and the scattering along the track. The length of the track is determined by the stopping power of the emulsion which depends on the proportions of silver halide and gelatin present. The grain spacing for lower energy particles, for example, protons <10 million electron volts, is dependent on the degree of development. In such instances a more moderate development may give a better evaluation of the grain spacing. With thick coatings, that is, 50 microns to 500 microns, the problem of depth development has been satisfactorily attacked, by retarding the action of the developer by lowering of temperature and by dilution until complete emulsion penetration has been realized. Preswelling the emulsion with water before processing also has proved an aid in depth processing. Fading of the latent image is an important problem in nuclear particle emulsion. Physical conditions after exposure influence the effect and in general low temperature and low relative humidity reduce or eliminate latent image loss.

Plates containing light elements are available, for example, boron, beryllium, lithium. These may be used for the study of nuclear reactions and for neutron monitoring.

Cloud Chambers; *G. C. Baldwin (General Electric Company, Schenectady, N. Y.).*

The three physical phenomena involved in the formation of a cloud track are, namely, the decrease in temperature of a gas which results from a sudden expansion of its volume, the condensation of moisture from a supersaturated vapor to form drops, the production of ions in a gas by a fast particle due to interaction of the field of its charge with the atomic electrons in the gas.

The chamber used with the 100 million electron volt betatron in Schenectady is 12 inches in diameter and employs coils supplying a 4,000-gauss magnetic field. This is a piston-type chamber; the gas-vapor mixture being contained in a glass enclosure, the base of which is a movable piston. The piston is raised periodically to compress the gas; by means of an electronically-controlled tripping device it is caused to drop suddenly at the proper time of each cycle. The resulting tracks are illuminated by an intense short-duration light flash and registed photographically.

Exact timing of these operations, so that the cloud chamber and associated equipment operate in synchronism with the action of the betatron is obtained by means of thyratrons triggered through *RD* delay networks by signals from a synchronizing circuit. Other operations such as pulsing the current in the magnet coils, winding film in the camera, are initiated by means of .cams. Ions are swept out of the chamber between expansions by means of an electric field applied between the piston and a transparent conducting coating on the inner surface of the upper plate of the chamber. A camera of special design with associated mirrors photographs the tracks so that a stereoscopic view is obtained.

Papers Digested for Conference on High-Frequency Measurements

These are authors' digests of most of the papers presented at the conference on high-frequency measurements, sponsored jointly by the AIEE, the Institute of Radio Engineers, and the National Bureau of Standards, Washington, D. C., January 10–12, 1949. The papers are not scheduled for publication in AIEE TRANSACTIONS or AIEE PROCEEDINGS nor are they available from the Institute.

Microwave Spectroscopic Frequency and Time Standards; *Harold Lyons (National Bureau of Standards, Washington, D. C.).*

The present primary standard of time is the mean solar day which slowly is growing longer, and fluctuates by about one part in 25 million. These and other limitations in astronomical standards make it desirable to look for new invariant and reproducible standards of time and frequency such as those using spectrum lines which also would improve world-wide frequency standardization now limited in accuracy by transmission effects in the ionosphere.

Spectrum lines of atoms or molecules in field-free regions and having very high natural Q would serve as ideal standards giving a more basic time unit than the arbitrary one provided by the rotation of the earth. The Q of spectrum lines is limited in practice by collision, Doppler, and saturation broadening. An inspection of the formulas for these effects indicates the desirability of using as high a frequency as possible in order to obtain high Q and high absorption coefficients, and also heavy molecules, low temperatures, and large absorption cells.

In the case of the inversion spectrum of ammonia, Q's of the order of 300,000 possibly might be attained. By going over to atomic beam techniques, collision and Doppler broadening can be eliminated. For a transition path length of 50 centimeters a Q of about 9 million could be obtained for a beam of cesium atoms and about 30 million for thallium.

The atomic clock using the 3,3 line of ammonia at 23,870.1 megacycles, which was developed at the National Bureau of Standards, is of the servo or frequency discriminator type. It is essentially a quartz-crystal clock which is regulated or locked by the ammonia line through a frequency-multiplier chain and servo control circuit. Frequency stability tests of the clock showed a constancy of better than one part in 20 million. The short time stability is provided by the quartz-crystal oscillator while the ammonia line determines the long time stability.

Another method of frequency control, also applicable to atomic clocks, was proposed in which absorption lines were used in circuits analogous to low-frequency quartz-crystal oscillators. These methods yield atomic oscillators or absorption-line oscillators independent of servo methods. By frequency divider circuits of the regenerative modulator or locked oscillator type the frequency could be reduced to the usual clock frequencies. The atomic oscillators are of the feedback type in which regenerative feedback to a microwave amplifier is controlled by an absorption cell. One such oscillator which was tested in an equivalent circuit design transmits the feedback signal through a magic-tee in which the cell terminates one impedance arm. Another oscillator uses a 6-arm junction as a waveguide Wheatstone bridge with the absorption cell used as one of the bridge terminations. A precision, absorption-line oscillator suitable as a frequency and time standard could be made in this way largely independent of external parameters. The lines of deuterated ammonia are being investigated for use in these oscillators.

An atomic beam clock of the servo-type, using a locked quartz-crystal oscillator and multiplier chain, offers the greatest potential for ultimate accuracy as a primary frequency and time standard. A potential accuracy of one part in ten billion or greater is indicated.

Frequency Stabilization With Microwave Spectral Lines; *W. D. Hershberger (RCA Laboratories, Princeton, N. J.).*

In the microwave frequency stabilization system,[1] a sweeping oscillator is employed to compare the frequency of a molecular spectral line f_m, and the operating frequency of a klystron f_k, and the relationship between the two frequencies is given by

$$f_k = \frac{f_m \pm f_e}{n}$$

where n is a whole number, and f_e is a relatively low off-set frequency or zero. As a result of the comparison an error voltage is developed and used to stabilize frequency.

The stabilizer acts degeneratively on frequency modulation impressed on the klystron over a frequency range limited ultimately by the band width of the spectral line, since the effective time constant of the filter in the servoloop is reduced by the stabilization factor $(1+\mu)$. This factor in the present system has been increased to well over 10^6. If Δf_0 represents the frequency change experienced by an unstabilized klystron due to an impressed disturbance, the stabilizer serves to reduce the change actually experienced by a stabilized klystron subjected to a similar disturbance to

$$\Delta f_0' = \frac{1}{1+\mu} \Delta f_0$$

Thus the short-term frequency stability depends both on the magnitude of an impressed disturbance and on loop gain μ.

The chief merit of the present stabilizer resides not so much in its high short-term stability realized by high loop gain as in its long term stability. Stability of this variety depends on how well and with what degree of permanence a set-point or reference frequency is established, namely, that frequency for which discriminator output changes sign. Uncertainties or drift in set point frequency are reproduced in the output frequency with fidelity given by the factor $\frac{\mu}{1+\mu}$. Thus if set-point frequency is changed by the amount Δf_s, for example by introducing a change in f_e, the output frequency change is given by

$$\Delta f'_s = \frac{\mu}{1+\mu} \Delta f_s$$

The two methods of introducing changes of frequency may be combined to accomplish wide band frequency modulation.

The ultimate basis of set-point stability is the molecular spectral line with its Q of 10^5 or higher. In the present state of the art, one need not differentiate between "molecular" or kinematic time and "dynamical" time.[2] However, it is not unlikely that data from molecular clocks will play an important role in testing current hypotheses on the origin of the red shift of astronomy.

Two stabilizers are now in use in experiments on short-term stability. Also some experimental work has been conducted on sealed off wave guide cells to hold ammonia at reduced pressure. Measurements on long-term stability will obviously require long-term frequency comparisons. An analysis of sources of error indicates that with an initial line Q of 10^5, long-term stability of 1 part in 10^7 is realized. This means that the servo loop is required to hold in time coincidence two wave forms arising from resonance phenomena to within $\frac{1}{100}$ the width of either. Further progress with systems such as the present one, in which a frequency comparison is made on the basis of time or phase, will depend on close attention to the characteristics of the method used in converting frequency to time or phase, to the constancy of amplifier characteristics, and to the differences in shape between resonance curves characterizing molecular spectral lines and those characterizing tuned electric circuits.

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The Stabilization of Microwave Oscillators; *E. W. Fletcher (Cruft Laboratory, Harvard University, Cambridge, Mass.).*

The problem of stabilizing a microwave oscillator using a cavity resonator as a reference standard was recognized and developed by R. V. Pound at the Massachusetts Institute of Technology radiation laboratory. Pound used a high-Q cavity as a frequency control element for a klystron oscillator by making a frequency comparison between the oscillator and the reference cavity.

Pound first used a direct-coupled amplifier in the feedback loop, and later, to overcome difficulties with drift and low-frequency noise, he developed an intermediate-frequency system at 30 megacycles per second.

One problem to consider is that of the reduction of noise in the klystron and its associated power supply circuits. This noise stems from three principal sources: shot noise, power supply instability, and variations in ambient conditions.

An analysis treating the stabilization system from a conventional servo point of view indicates that an improvement in stability may be had with a higher-Q reference standard. In a study of the problem of high-Q reference standards, including both cavity resonators and molecular absorption lines, it is shown that the latter has distinct advantages as a reference standard.

This research was made possible through support extended Cruft Laboratory, Harvard University, jointly by the Navy Department (Office of Naval Research), the Signal Corps of the United States Army, and the United States Air Force, under ONR Contract N5ori-76, T. O. I.

Superconducting Resonant Cavities—Measurements of the Surface Impedance of Normal and Superconductors at Low Temperatures and Microwave Frequencies;* *E. Maxwell (Massachusetts Institute of Technology, Research Laboratory of Electronics, Cambridge, Mass.; now with the National Bureau of Standards, Washington, D. C.).*

Measurements have been made on the properties of resonant tin cavities at temperatures down to two degrees Kelvin and at a frequency of 24,000 megacycles per second. Interest centers both in the behavior in the normal state at temperatures of 30 degrees Kelvin, and below as well as in the superconducting behavior. In the normal state it is found that the cavity Q does not rise proportionately as the d-c conductivity is increased by lowering the temperature but rather reaches a limiting value which is independent of the temperature after about 15 degrees Kelvin. The limiting value is a characteristic of the metal and the frequency. This anomalous behavior is due to the fact that the electronic mean free path becomes large compared to the skin depth and consequently the simple conductivity concept becomes inadequate at these temperatures. Instead one uses the surface impedance concept, which defines the conducting property of the metal in terms of the ratio of tangential E to tangential H at the surface of the metal.

A large and relatively sudden increase in the Q is observed when the temperature of the cavity is lowered beyond the superconducting transition temperature. The transition is not as sharp as observed in the case of direct current nor does the resistance vanish completely as for the d-c case. Q's of the order of a few hundred thousand to a few million may be obtained.

The measurement techniques center on methods of measuring high Q's. Several schemes, separately and in combination, have been used. In some cases resonance curves were taken using pound stabilized continuous-wave oscillators and either a slotted section or magic tee for measuring input standing wave ratio as a function of frequency. For rapid measurements a scheme using frequency-modulated klystron together with a magic tee, crystal detector and oscilloscope was employed to give visual display of the resonance absorption of the cavity. This arrangement permits rapid measurement of the standing wave ratio at resonance and, hence the ratio of unloaded to window Q, as a function of temperature.

* This work was supported by the Army Signal Corps, Navy Department, and Air Force under Signal Corps Contract W-36-039-SC 32037.

Sources of Power for Microwave Measurements; *George E. Hackley (Sperry Gyroscope Company, Great Neck, N. Y.).*

The characteristics required of the power source for microwave measurements are quite varied and they are determined by the sensitivity of the measuring equipment and by the measurement being made. For most purposes a power output of about one watt is adequate, although some special measurements, particularly voltage breakdown tests, require very large powers. Good frequency stability is quite important in measurements of narrow-band or resonant circuits, while wide tuning ranges are required for broadband testing of components. The source should be easy to frequency modulate and amplitude modulate. To obtain all of these varied characteristics, there are a number of different types of sources of power, although no single type fulfills all of the requirements.

Triodes may be used as amplifiers, multipliers, or oscillators at frequencies up to about 3,000 megacycles with power outputs up to 25 watts. They are not easily frequency-modulated, but they may be pulse-modulated easily to obtain very high values of peak power. Triodes are most useful as power sources at frequencies below about 2,000 megacycles.

Magnetrons, which are oscillators, may be used from a few hundred megacycles to about 50,000 megacycles, although they are not available for every frequency within this range. Most magnetrons are fixed tuned and must be pulsed, but a few continuous wave and narrow tuning range ones are available. Frequency-modulated magnetrons have been designed, but they are not generally available. Peak powers of kilowatts or megawatts may be obtained from magnetrons. As a power source for microwave measurements, magnetrons are most useful for high-power testing of components.

Crystal rectifiers may be used as harmonic generators to provide very small amounts of power at frequencies up to 50,000 megacycles or more. Because of the small amount of power that may be obtained, they are most useful at frequencies where other sources of power are not available or where only small powers are needed, as in frequency standards.

Traveling wave tubes, which are in the experimental stage at present, are wide-band amplifiers which may prove to be quite useful as power sources. They may be used as oscillators, by providing feedback through a resonant circuit, over fairly wide ranges with power outputs in the order of one watt at a few thousand megacycles.

Klystrons, which are available as amplifiers, multipliers, and oscillators, are the most useful of the sources of power for microwave measurements and they may be used at frequencies up to 50,000 megacycles. Klystron amplifiers and multipliers are rather narrow-band devices and they are most useful for fixed frequency applications, as in frequency standards. Power outputs of over 100 watts may be obtained from klystron amplifiers at 3,000 megacycles and correspondingly less power may be obtained at higher frequencies. Reflex klystrons, which are oscillators with only one resonant circuit, are the most useful and available source of microwave power. Power outputs of reflex klystrons range from over 10 watts at 2,000 megacycles to more than 5 milliwatts at 50,000 megacycles. Reflex klystrons are tuned easily and may be frequency, pulse, or square-wave modulated.

Bolometric Measurement of Microwave Power Over Broad Frequency Bands;* *Herbert J. Carlin (Microwave Research Institute, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.).*

The most usual practice in the application of bolometers for the measurement of microwave power is to use a substitution procedure which equates low-frequency power to microwave power by means of a Wheatstone bridge. For this method to be valid, it is desirable that the following general design criteria be satisfied:

1. The bolometer cross-sectional dimensions should be less than the skin depth to insure similar radio-frequency and bias power distributions over the bolometer cross section.

2. The bolometer should be short compared to wavelength to insure similar radio-frequency and bias power distribution along the bolometer length.

3. The bolometer should be long compared to cross-sectional dimensions in order that the thermal properties be uniform along the bolometer length.

4. The static resistance-power curve of the bolometer should be linear. If this, as well as number 3, is closely satisfied, then the substitution method is valid independent of the longitudinal radio-frequency versus low-frequency power distribution. In addition, linearity minimizes some of the errors due to a pulse modulated radio-frequency signal.

5. The bolometer should have a relatively long thermal time constant. If not, the bolometer resistance follows a pulsed signal and the null detector of the substitution bridge circuit has a varying component of current. This leads to an incorrect bridge balance and a resulting error under pulsed conditions.

It is desirable that a bolometer power meter operate over as broad a frequency band as possible with a fixed tuning adjustment. The Microwave Research Institute of the Polytechnic Institute of Brooklyn has developed a series of 6 power meters which cover the frequency band 20–10,000 megacycles per second, and a power range of 25 microwatts to 5 watts. The maximum voltage standing wave ratio is 1.30, representing a maximum of one per cent reflection loss with a matched generator. Extremely broad-band performance is obtained by means of an adjustable short-circuiting slug at the back of the bolometer, and an adjustable undercut section of line in front of the bolometer. These tuning devices cancel the parasitic reactance associated with the bolometer. They are set experimentally for optimum voltage standing wave ratio over the desired frequency band and then locked permanently in position.

* This paper represents research done at the Microwave Research Institute of the Polytechnic Institute of Brooklyn sponsored by the Watson Laboratories, AMC, United States Air Force, under contract W-33-038ac-13848.

Microwave Metallized-Glass Attenuators;* *J. W. E. Griemsmann (Microwave Research Institute, Polytechnic Institute of Brooklyn, Brooklyn, N. Y.).*

In addition to the known applications of microwave attenuators in signal generators and as multipliers for power meter, attenuators occupy a place of considerable importance in the whole field of microwave instrumentation.

Both wave guide and coaxial attentuators have been developed for specific broad bands of frequencies in the over-all range from 0 to 25,000 megacycles per second. A set value of attentuation is dependent in only small measure upon the frequency, a typical band for the coaxial fixed value attenuator being 1,000 to 4,000 megacycles per second. The input and output impedances are matched in any case to a voltage standing wave ratio value of less than 1.3. In the case of the variable waveguide attentuators, the low minimum insertion loss (<0.5 decibel) increases greatly the usefulness of these instruments for measurements.

All of the attenuator developments use, as the basic dissipating element, a very thin resistive metallic film formed on glass. Films of noble metals alloys or nichrome are used.

Coaxial fixed pads in the attenuation values 3, 6, 10, and 20 decibels have been developed for the frequency range 4,000 to 10,000 megacycles per second, 1,000 to 4,000 megacycles per second and 0 to 1,000 megacycles per second.[1] The designs of the coaxial units are based on distributed parameter transmission line theory and are constructed on the basis that the resistance per square measured at power or audio frequencies is identical to that active at the microwave frequencies. Special equivalent circuit techniques were derived for frequencies below 4,000 megacycles per second which allow analytical design of the extremely broadband "chimney" types of attenuators. Departure from the distributed parameter transmission line theory occurs when the attenuation rate becomes very high.

Continuously variable waveguide attenuators have been developed with satisfactory performance over the full useful range of the waveguide. The following table gives the sizes of units developed and their corresponding performance characteristics.

Guide Size (In.)	Frequency Range (kmc/sec)	Maximum Insertion Loss (Decibel)	Attenuation Spread at 40-Decibel Attenuation
$1/2 \times 1/4 \times 0.040$	18.0 to 25.6	0.5	±2.5
$1 \times 1/2 \times 0.050$	8.3 to 12.15	0.5	±1.3
$2 \times 1 \times 0.057$	3.95 to 5.85	0.5	±1.5

For the vast majority of settings the voltage standing wave guide is less than 1.1, but, in any case, the voltage standing waveguide never exceeds 1.2.

Development on the $1\frac{1}{4}$ inch by $\frac{5}{8}$ inch guide size is nearly complete and that on the 3 inch by $1\frac{1}{2}$ inch started.

The first attenuator in the list is of the guillotine type in which a resistance film formed on a thin insulating glass base is introduced through a slot in the top or broad side of the waveguide. The performance, particularly the attenuation spread, is critically dependent on those factors which control the feed of current to the resistance film. Better performance is obtained for very thin insulating base, such as the 0.007-inch mil glass plate used in the foregoing attenuator.

The remaining two attenuators are the vane type which uses a flat metallic resistive plate with surface parallel to the electric field lines. Variability is attained by moving the plate from the side of the waveguide into the more intense fields at the center. The performance which can include such occurrences as a "resonance-like" attenuation curve are dependent critically on the plate width and resistivity again as a result of the method of feeding currents to the resistance film.

Other special broadbanding applications for the resistance films are anticipated.

* The developments described have been sponsored by the Navy Bureau of Ships and Watson Laboratories, AMC, United States Air Force under contracts with Polytechnic Institute of Brooklyn.

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1. A Bead-Supported Coaxial Attenuator for the Frequency Band 4,000–10,000 Megacycles per Second, **H. J. Carlin, John W. E. Griemsmann.** *NEC Proceedings*, volume 3, 1947, page 79.

The Effective Conductivity of Wires at Microwave Frequencies; *A. C. Beck, R. W. Dawson (Bell Telephone Laboratories, Holmdel, N. J.).*

When studying the attenuation of coaxial lines and waveguides at microwave frequencies, it is desirable to know the heat loss in the metal itself for different materials and surface conditions. To obtain this information, equipment was set up to determine the effective conductivity of small wire samples at frequencies in the order of 9,000 megacycles.

A short piece of the sample to be tested is supported by polyfoam beads in an open-ended tube somewhat longer than the sample. The resulting open-circuited coaxial line is excited through an opening from a terminated waveguide by a signal oscillator which is frequency-modulated about a resonant frequency of the line. The signal taken into a second waveguide through an opposite opening in the coaxial line is connected to a superheterodyne receiver whose beating oscillator is frequency-modulated in synchronism with the signal oscillator. The output of this receiver is the resonance curve of the test line. A cathode-ray oscilloscope is switched at a 30-cycle rate between this output and the trace of the rectified envelope of the signal oscillator on which a frequency marker from a tunable resonant cavity wavemeter is superimposed. By using an accurately calibrated attenuator to determine the levels, the half-power bandwidth of this resonance curve is measured with the cavity wavemeter. The Q is determined by dividing the frequency at resonance by this bandwidth. It is corrected for coupling losses by measuring the transmission loss through the specimen holder at resonance. Corrections are made also for outer conductor losses and end effects. As a result, the absolute effective conductivity of the wire sample is obtained.

Measurements on commercial wires of high conductivity, such as silver, copper, gold, and aluminum, give Q values about 85 to 90 per cent of those calculated from the measured d-c conductivity. Wires of lower conductivity, such as molybdenum, platinum, brass, and phosphor bronze, give values from 90 to 100 per cent of the calculated value. Very smooth electropolished copper wires measure close to 100 per cent, but drop to about 95 per cent if left exposed in the laboratory unless lacquered. This indicates the importance of a smooth surface, especially for high conductivity samples, at these frequencies where the current skin depth is so exceedingly small. Smooth copper wires rubbed lengthwise with fine emery cloth dropped to about 80 per cent of their initial value, and rolled between ground steel plates to produce circumferential grooves about three microns deep dropped to about 60 per cent of their initial value. A copper wire with 220 threads per inch had only 40 per cent of its initial Q. Copper electroplated on very smooth electropolished copper wires gave values about 80 per cent of the initial value, and when re-electropolished the value was only 95 per cent of the original value. Silver electroplated on very smooth electropolished copper wires from a cyanide bath at various rates gave Q values from 45 to 75 per cent of the initial value, the higher results for the smoother platings. Only a few plated samples have been measured so far, and this work is being continued.

As a check on the accuracy of the equipment, it was found that very smooth samples of platinum, phosphor bronze, brass, molybdenum, 24 carat gold, and copper gave results within two per cent of the Q value calculated from their measured d-c conductivity. These metals have Q's spread over the range from 800 to 2,100 when used in this specimen holder.

The information obtained with this measuring equipment has been found useful in the design and treatment of microwave components of all types where loss is a factor of importance.

X-Band Phase Shiftless Power Splitter; *Henry J. Riblet (Submarine Signal Company, Boston, Mass.).*

For some time, there has been a recognized need for an item of X-band test equipment which will act as a continuous power divider without introducing any relative phase shift between the output voltages. A device which will do this for a 30-decibel range in power division, at all frequencies in the X-band, within a tolerance of about ± 1 degree, consists of a directional coupler of the perpendicular slot variety[1] in which the common wall between the wave guides may be moved laterally between the guides. The trick depends on the fact that by placing the longitudinal slots on only one side of the guide, it is possible to vary the plate from the position of maximum power transfer to the position of minimum power transfer without encountering any current discontinuities.

In the experimental model, a sufficient number of slots were used so that it behaved essentially as a hybrid when adjusted for maximum power transfer. For minimum power transfer the insertion loss was about 35 decibels or 0.1 decibel depending on which of the output terminals was investigated. Thus, operating simply as an attenuator, the device covers the attenuation range from 0.1 decibel to 35 decibels.

The fact that it splits power without introducing any relative phase shift between the output voltages depends on its symmetry as lossless directional coupler. If φ is the relative phase between the output voltages, if δ is its directivity, and if ρ is the magnitude of its reflection coefficient, it is a rather simple matter to show that

$$|\cos \varphi| \leq \epsilon\sqrt{\delta}\rho,$$

where ϵ varies from 1.4 for hybrid performance to 1 for no power transfer. Over most of the range of operation of this device there is no difficulty in obtaining directivities > 20 decibels and standing wave ratios $\leqslant 1$ decibel. Hence $\sqrt{\delta} \leqslant 0.1$ and $\rho \leqslant 0.06$. Thus $|\cos \varphi| \leqslant 0.01$ and 89 degrees $< |\varphi| < 91$ degrees.

The experimental results confirm this estimate in that, for five frequencies covering the 12 per cent X-band and for all attenuations ranging from 0 decibel to 30 decibels, the relative output phases are constant to within ± 1 degree. Slight deviations of this constant from 90 degrees may be attributed to mechanical assymetries.

In addition to its use as a phase shiftless power divider, it serves quite nicely as a directional coupler and very stable calibrated attenuator. In this connection we have used it extensively as a medium power attenuator for duplexer measurements.

REFERENCE

1. A New Type of Waveguide Directional Coupler, **H. J. Riblet, T. S. Saad.** *Proceedings*, Institute of Radio Engineers (New York, N. Y.), volume 36, January 1948, pages 61–4.

A Figure of Merit for Directional Couplers; *George James (Sperry Gyroscope Company, Great Neck, N. Y.).*

Two definitions of the concept of directivity appear in the literature. One of these, intrinsic directivity, measures the properties of the coupling structure only and is limited in significance except to the designer of directional couplers. The second, effective directivity, takes account also of the properties of the secondary line termination, has the properties of a figure of merit and is readily measured.

In order to display the properties of the concept of effective directivity we consider a generic coupler in which the primary line amplitude is assumed to be unity. Secondary line amplitudes are taken to be A_1 and A_2 in the high and low level ends respectively. The primary line load is assigned the coefficient of reflection r_1 and r_2 is the coefficient of reflection of the secondary line termination.

Using this terminology we define:

1. Coupling: $C=20 \log \dfrac{1}{A_1}=-20 \log A_1$.

2. Directivity coefficient: $d=A_2/A_1$.

3. Intrinsic directivity:

$$D=20 \log A_1/A_2=20 \log \frac{1}{d}$$

4. Effective directivity:

$$D_{eff}=20 \log \frac{1}{|de^{j\varphi i}+r_2e^{j\varphi ii}|}$$

Where A is total resultant amplitude incident on M and φi and φii are phase angles determined by the electrical lengths of the paths of propagation within the coupler, by the phase angle of r_2, and by the phase shifts in the coupling device.

In measurement of power we consider first the effects of the parameters r_1, r_2, and d upon the accuracy of power measurement with a directional coupler. It may be shown that the amplitude of the wave incident upon the response device M is

$$A_1|1+r_2de^{j\varphi iii}+r_1e^{j\varphi iv}(de^{j\varphi ii}+r_2e^{j\varphi ii})| \qquad (1)$$

where all terms of order three or greater in r_1, r_2, d have been ignored. Since r_2 and d are inherent properties of the coupler and since the paths of propagation which affect φ_{iii} do not involve reflection from r_1. the term $r_2de^{j\varphi iii}$ may be and usually is included in the calibration of the instrument. The measurement of relative power level involves an error in amplitude determined by the phase difference between the vectors

$$1+r_2de^{j\varphi iii} \qquad (2)$$

and

$$r_1e^{j\varphi i_v}(de^{j\varphi i}+r_2e^{j\varphi ii}) \qquad (3)$$

and will be greatest for phase differences of either 0 or 180 degrees between equations 2 and 3. Thus the maximum error in power measurement is approximated by

$$r_1/de^{j\varphi i}+r_2e^{j\varphi iii}/ \qquad (4)$$

It may be shown also that the error in measurement of coefficient of reflection using a directional coupler is approximated by

$$/de^{j\varphi i}+r_2e^{j\varphi iii}/ \qquad (5)$$

Equations 4 and 5 each may be represented in simple graphical form.

A Precise Direct Reading Phase and Transmission Measuring System For Video Frequencies; *D. A. Alsberg, D. Leed (Bell Telephone Laboratories, Inc., Murray Hill, N. J.).*

An insertion phase and transmission measuring system has been developed which combines laboratory precision of measurement with speed of operation suitable for use in production testing. The system covers a frequency range of 50 to 3,600 kc. Its accuracy is in excess of ±0.25 degree in phase over a continuous range of 360 degrees, and ±0.05 decibel in transmission over a range from 60-decibel loss to 40-decibel gain. A heterodyne measurement method is used.

Reactance tubes automatically control the slave oscillator frequency from the outputs of a pair of discriminators, a frequency discriminator with an output proportional to frequency error, and a phase discriminator with an output proportional to the integral of frequency error. This combination maintains the slave oscillator at 31-kc difference with respect to the master oscillator without sideband ambiguity. The d-c loop gain is effectively infinite and the frequency error therefore zero. The frequency discriminator insures noncritical lock-in.

The circuit accuracy is limited by the range over which a frequency converter responds linearly to input voltage changes. A *6AK5* vacuum tube is used as a square-law-type converter. A linearity of 0.01 decibel is obtained over an input voltage range of more than 30 decibels.

An equal 4-arm phase bridge with two outputs is used as the phase sensitive element in the detector. Equality between these outputs defines the null of the system regardless of the input voltage amplitudes to the bridge. Amplitude difference between these outputs measures directly phase difference between the input voltages, independently, to a first order, of input amplitude inequalities.

The phase shifter employs a 4-quadrant variable sine capacitor. The residual error of the phase shifter is corrected automatically by an optically projected moving index. Linear scales may be set to establish an arbitrary zero reading. Unique balance indications of the phase bridge occur every 360 degrees and define an absolute standard of 360 degree phase shift. Exact submultiples of 360 degrees are generated from this standard by the method of substitution and used to produce the optical corrector. The phase shifter is calibrated to an absolute accuracy of better than ±0.1 degree. Higher accuracy could be obtained with the calibration circuit if the need arose.

By means of a simple automatic dial lighting system the measuring attenuator indicates directly the gain or loss of apparatus under test.

The following are the main features.

The measurement of phase is unambiguous with respect to quadrants and the measurements of insertion phase and loss or gain are independent of each other. The entire frequency range is covered without band switching by use of a heterodyne signal oscillator, and the system zero is independent of measurement frequency. Detector tuning is eliminated through the use of frequency conversion, employing a beating oscillator automatically controlled in frequency by the signal oscillator. Phase and transmission may be read directly, without auxiliary computations, from the dials of the phase shifter and attenuator or the scales of indicators.

Methods of Measuring Impedance and Voltage Standing-Wave Ratio at Microwave Frequencies; *F. Klawsnik (Sperry Gyroscope Company, Great Neck, N. Y.).*

The following methods may be considered.

Standing Wave Detector. That deep probe penetration, makes it possible to obtain exact results even from badly distorted patterns is illustrated. A variation is the use of line stretchers on one or both sides of a fixed probe.

4-Probe Sampling of Electric Fields. The two probes of each pair are separated by a quarter wave length and the pairs are staggered by an eighth wave. The output from each probe goes to a crystal detector. The difference in the output of each probe of one pair ($2V_iV_R \cos\theta$) is impressed on the horizontal plates of a scope and the difference in the output of each probe of the other pair ($2V_iV_R \sin\theta$) is impressed on the vertical plates. When the microwave source is frequency-modulated, a trace of the reflection coefficient will appear on the scope. With adjustable probes the impedance can be read to better than five per cent for the full bandwidth.

2-Slot Sampling of the Electromagnetic Fields. Two cross guides, one either side of the main guide, are coupled to the main guide through a slot. One slot is excited by the total longitudinal current flow and is proportional to the vector difference of the incident and reflected voltages. The other slot is excited by the total transverse current and is proportional to the vector sum of the incident and reflected voltages. The input to the main guide is frequency-modulated. The output from the two cross guides is $V_i+V_R+2V_iV_R \cos\phi$ and $V_i+V_R-2V_iV_R \cos\phi$. The ratio of these outputs is applied to the vertical plates of a scope and the horizontal plates are connected to the sweep frequency. The trace on the scope will represent the square of the voltage standing wave ratio. This method can be used over a 12 per cent frequency range and gives an accuracy of 8 per cent in magnitude and 4 per cent in phase.

Microwave Bridges. Two types of bridges are the 6-arm exact equivalent of the Wheatstone bridge, and the hybrid circuit used as a bridge. In the 6-arm bridge the voltage standing wave ratio is determined by three physical lengths. Here it is possible to measure any impedance with about the same accuracy as with a standing wave detector.

Most hybrid circuit bridges are not balanced to a null but are based on the fact that the output is proportional to the square of the reflection coefficient of the test piece when one arm is terminated in a matched load. An indicator on the output arm can be calibrated in voltage standing wave ratio. For badly mismatched test pieces a movable short circuit may be used instead of the matched load and then the output is a measure of one minus the reflection coefficient.

The reflectometer is an extension to directional couples of the hybrid circuit bridge for measuring the voltage standing wave ratio.

Resonance Cavity. The length of a resonant cavity is determined by the magnitude and position of a discontinuity in this cavity. With adjustable shorts it is possible to form a cavity and to move it relative to a discontinuity on the line.

The varying length of the cavity is a measure of the magnitude of the reflection coefficient. If proper precautions are taken it is possible to measure both extremely low voltage standing wave ratios in the order of 1.002 to an accuracy of one-tenth of one per cent and also to measure extremely high voltage standing wave ratios in the order of 7,000 to 1 to an accuracy better than one per cent.

This method is used to measure nondissipative elements such as slot couplings, windows, steps, connectors, and posts to very extreme accuracies.

Generator Mismatch Measurements in Transmission Lines; *Peter E. Gilmer (Bell Telephone Laboratories, New York, N. Y.).*

The mismatch "looking back" along a transmission line towards an active signal generator or source can be measured by the following simple method. This method is valuable because a mismatched generator can cause signal distortions in communications systems and errors in measurements.

The interaction, or repeated reflection effects, that are responsible for the distortions and errors just mentioned, are utilized in the measurement. The method amounts to varying the phase between the generator and a purposely mismatched load. Multiple reflections between the generator and the load cause the output to change as the phase is changed. The generator mismatch can be determined by proper interpretation of the measured output variations.

A device that can be used to measure the generator mismatch in rectangular waveguide consists of the following. A coupling loop projects through a movable short-circuiting piston and connects to a coaxial plug attached to the back of the piston. In this particular device a crystal detector is built into the center conductor for use with a single detection circuit or a d-c milliammeter. The procedure is as follows:

1. Insert the piston into a length of waveguide so that the piston "looks back into" the generator mismatch to be measured. The generator must be turned on.

2. Connect an indicating circuit to the piston output connection.

3. Move the piston over a range of at least a half wave length in the guide.

4. The output will be observed to vary in much the same manner as if a standing wave detector were being used in a conventional measuring circuit. Note the ratio of maximum to minimum current picked up by the loop, S_a, an apparent standing wave ratio. Account must be taken of the detector characteristic.

(*a*). For an approximate solution, the generator standing wave ratio, S_g, is approximately equal to the apparent standing wave ratio, S_a. This assumes that the reflection coefficient of the piston, ρ_p, is almost unity and is large compared to the reflection coefficient of the generator, ρ_g.

(*b*). For an exact solution, use the following equation. The reflection coefficient of the piston, which must be known for this case, can be measured by conventional means:

$$\rho_g\rho_p = \frac{S_a-1}{S_a+1}$$

The reflection coefficient of the piston should be chosen keeping the following limitations in mind. A small reflection coefficient reduces the sensitivity of the measurement. In some instances, a large reflection coefficient is undesirable. The varying load presented to the generator by the movement of the piston may pull the frequency or output of the generator.

This method can be used also to measure the mismatch of a 4-terminal network if the generator impedance is first matched to the transmission line and then the unknown network connected in tandem with the generator.

If it is desired to use this method for measuring very small mismatches, the piston must be designed carefully and it must be used in a precision section of wave guide so as to keep the reflection coefficient of the piston constant.

A Method of Measuring Phase at Microwave Frequencies; *Sloan D. Robertson (Bell Telephone Laboratories, Inc., Holmdel, N. J.).*

A method of phase measurement using a homodyne detection principle which operates in the following manner. The output of a microwave signal oscillator is divided in two portions. One portion is applied to a balanced modulator where it is modulated by an audio-frequency signal. The suppressed-carrier double-sideband signal from the modulator is applied to the device whose phase shift is to be measured. Means are available for sampling the signal at both the input and output of the device. The other portion of the oscillator power is fed through a calibrated phase shifter and is applied to a crystal detector in the manner of a local oscillator of a double-detection receiver. The signal samples then are applied alternately to the crystal detector where they are demodulated by the action of the homodyne carrier. In each case the phase shifter is adjusted so that the audio signal disappears in the detector output. This occurs when the phase of the homodyne carrier is in quadrature with the signal sidebands. The difference in phase between the two adjustments of the phase shifter is equal to the phase difference between the two samples. The apparatus can be assembled with standard waveguide components.

Dielectric Measurement Techniques in the Very-High-Frequency Region;† *W. B. Westphal (Laboratory for Insulation Research, Massachusetts Institute of Technology, Cambridge, Mass.).*

High-frequency measurements of dielectric properties are based on the interaction between the material and electromagnetic waves. The dielectric constant (ϵ^*) and the magnetic permeability (μ^*) are complex numbers expressing the ratio of flux density to field intensity for the electric (E) and magnetic (H) fields respectively. The ratio of energy dissipated to energy stored defines the loss tangent for material:

$$\epsilon^* = \epsilon' - j\epsilon'' \quad \tan\delta_d = \frac{\epsilon''}{\epsilon'}$$

$$\mu^* = \mu' j \mu'' \quad \tan\delta_M = \frac{\mu''}{\mu'}$$

For a transverse wave travelling in an infinite medium, the ratio E/H gives the intrinsic impedance $\left(\mathcal{Z} = \sqrt{\frac{\mu^*}{\epsilon^*}}\right)$ for the material. The progress of the wave is characterized by the complex propagation constant:

$$\gamma = j\frac{2\pi}{\lambda}\sqrt{\epsilon^*\mu^*} = \alpha + j\beta$$

In principle, the obvious measurements would be of amplitude at two points to determine α, a phase difference measurement at two points to determine β, and a complex impedance measurement at one point. If a normal reflecting boundary is added to the system, a standing wave pattern is produced which eliminates the necessity for instantaneous phase measurements. Instead, the distance between successive minima is a measure of β and the envelope decay characteristic determines α. In addition, an impedance measurement at one point is necessary. Alternately, impedance measurements at two points determine the desired quantities.

Measurements of either the E or H standing wave pattern in a known medium, terminated in a slab of unknown properties, determine the impedance at the boundary:

$$\mathcal{Z}_B = \mathcal{Z}_{01}\frac{\dfrac{E_{\min}}{E_{\max}} - j\tan\dfrac{2\pi X_0}{\lambda_1}}{1 - \dfrac{E_{\min}}{E_{\max}}\tan\dfrac{2\pi X_0}{\lambda_1}}$$

$\mathcal{Z}_{01}$ = the intrinsic impedance of the known medium

$\dfrac{E_{\min}}{E_{\max}}$ = the inverse standing wave ratio

X_0 = the distance between slab boundary and first minimum in the standing wave pattern

Two such impedance measurements, made with different lengths of sample slab, allow the unknown constants to be calculated. When the sample is known to be nonmagnetic, a single impedance measurement suffices.

Actually measurements are usually made in waveguides to overcome two inherent difficulties with space measurements:

1. The divergence of beams limits sensitivity for loss measurements.
2. Higher order modes introduce uncalculable error except in large sheets of samples.

Common methods of measurement of ϵ^* use a sample resting either against the short-circuited end of a transmission line, or a quarter wave length from it. For these two cases, curves of X_0 and $E_{\min}/E_{\max}$ were plotted as functions of wave length in the sample. From these graphs the rate of change of measured quantities with ϵ' and $\tan\delta$ can be obtained. With the rate of change known and the experimental error in measuring the probe displacements determined, the per cent error of the measurements can be calculated.

Resonance and standing wave methods are basically the same and can be shown to have identical accuracy for the same per cent error in measuring the incremental quantities.

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Measurement of the Electrical Characteristics of Quartz Crystal Units by Use of a Bridged Tee Null Network; *Charles H. Rothauge (The Johns Hopkins University, Baltimore, Md.).*

Specification of quartz crystal units has been made in the past by use of a test oscillator with the scale of activity the rectified grid current. However, it is possible to specify a crystal by use of its equivalent electrical constants. A bridged tee null network has been used to measure the equivalent resistance and the equivalent reactance of a crystal plate.

A tee configuration of capacitance and resistance is used which has a transfer impedance equivalent to a negative resistance and a capacitive reactance. The crystal operating at a frequency between series-resonance and antiresonance can be represented by a resistance in series with an inductive reactance. The tee bridged with a crystal will form a null transmission network, and the electrical constants of the crystal may be measured at its operating frequency.

This measuring circuit has the advantages that shielding is relatively simple (both the source and the detector have a common grounded terminal), and that corrections for all stray capacitances that affect the measurements may be included in the calibration of the capacitors of the tee. Disadvantages of the method are that coupling between the source and detector will affect balance conditions; also a variable resistor is required.

Buffer amplifiers of conventional design adequately alleviate the difficulty of coupling, and a variable resistance box was constructed that maintains its values of resistance within five per cent of its d-c values up to frequencies of seven megacycles per second with almost constant capacitance to its shield. This box gives values of resistance of 0 to 15,210 ohms in steps of ten ohms with a capacitance to its shield of 23.5 ±0.3 micromicrofarads for all values of resistance. The box was constructed in such a manner that at all times three resistance elements are in series. Thus with reasonable care in its construction capacitance to the shield is constant. The individual resistance elements are deposited carbon resistors which have relatively small reactance and constant effective resistance.

Measurements have been made on four series of crystals at frequencies ranging from 4.5 megacycles per second to about 7.5 megacycles per second. Values of equivalent reactance were determined from about 200 ohms to 750 ohms, and values of equivalent resistance varied from 10 ohms to 35 ohms.

The precision of these measurements is estimated to be ±0.3 per cent for the determination of the equivalent reactance and ±2.3 per cent for equivalent resistance.

Measurement of Artificial Dielectrics for Microwaves; *Winston E. Kock (Bell Telephone Laboratories, Murray Hill, N. J.).*

One of the important uses of dielectrics in high-frequency radio work is in the field of microwave lenses. In many applications, lenses have certain advantages over other microwave antennas, and a satisfactory dielectric or refracting material is thus desirable. Natural dielectrics, such as polystyrene, have the disadvantage of great weight, and to circumvent this difficulty, several types of lightweight artificial dielectrics have been developed at the Bell Laboratories.

The first refracting medium was made of thin conducting plates which acted as waveguides and caused radio waves passing through them to assume a higher phase velocity. The index of refraction was thus less than unity and a concave shape was necessary to produce a converging lens. Such lenses, ten feet square, are now in use in the Bell System microwave radio relay link between New York and Boston.

These first lenses, however, because of their waveguide properties, possessed bandwidth limitations, and the desirability of antennas capable of accommodating wider bands for use in the New York-Chicago radio relay route led to the development of a broad-band lightweight artificial dielectric. These "metallic" dielectrics comprised arrays of conducting elements such as spheres, disks, or strips, and the polarization of the conductors simulated the polarization and hence the refractive power of a true dielectric for wavelengths long compared to the size and spacing of the elements[1].

Early theory does not yield refractive indexes which agree well with experiment for practically desirable values of 1.5 or 1.6, so that the properties of the dielectrics must be investigated experimentally.

One such method is similar to the waveguide technique for measuring the properties of true dielectrics. Metal strips (or dielectric strips upon which squares or disks of conducting coating have been deposited) are inserted into a polystyrene foam support. From the position of the standing wave minimum and the magnitude of the standing wave, the dielectric constant and the loss tangent can be deduced.

A second method for evaluating the refractive index is to measure the optimum focal length of a complete lens; this may be done either by phase measurements across the aperture of the antenna[2], or by an examination of the directional patterns for various focal lengths. The optimum focal length yields the flattest phase fronts and the deepest nulls in the directional pattern. Upon observing directional patterns of a 7-centimeter microwave lens of 6-foot aperture with the feed placed at various distances behind the lens, the depth of the nulls shows the true focal length to be about 62 inches. Since a one inch axial departure of the feed horn from the true focus on this lens causes the phase front to deviate from a plane by only $\pm 1/60$ of a wave length, it can be concluded that this method is quite sensitive. A one inch change in the observed optimum focal length in this lens will change the evaluated index of refraction (which in this case is approximately 1.5) by 0.005.

One interesting observation is that the microwave properties of certain of the foregoing arrays can be inferred from acoustic measurements. The various obstacle array lenses were found to focus sound waves as well as microwaves so that measurement of the refractive index of certain structures, such as the strip array, could be made by the simpler acoustic measurements. Since 3.45-centimeter sound waves have a frequency of about 10,000 cycles, a continuous variation of wave length from one centimeter or less up to 10 or 20 is readily available acoustically.

A 6-inch square strip prism of open construction which was originally made for 3-centimeter microwave experiments was tested acoustically, and, from the angle of deviation which it imparted to the beam of a 6-inch horn, its refractive index could be ascertained for various frequencies. Such a curve was plotted. By spot checks, it was ascertained that, for like wave lengths, the measured acoustic index of refraction agreed with the measured microwave index.

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Microwave Noise Sources; *I. Mirman, J. H. Vogelman (Watson Laboratories, Red Bank, N. J.), R. H. George (Purdue University, Lafayette, Ind.).*

Pulsed-type noise generators are of either the coaxial or flat disk radial line type and utilize the discharge of a charged circuit as the source of noise. Units have a resonant frequency that is selected to give a desired noise spectrum output of constant amplitude over a frequency band up to at least 75 per cent or more of the resonant frequency. The output characteristic curve of the pulse-type generator resembles that of a low pass filter and proper selection of the peaking frequencies during design can be used to obtain any desired response characteristic. The coaxial noise generator consists of an open-ended coaxial line charged through a high resistance and discharged by short-circuiting inner and outer conductors. The contact between the inner and outer conductors is made by means of a driven inner conductor contacting a fixed flat point on the center conductor of the coupling transmission line which is at the same d-c potential as the outer conductor. The very short switching time produces a pulse of very narrow width and very sharp rise time with the resulting high-frequency components. Design relations predict possibility of reaching 375,000 megacycles per second utilizing flat disk contacts.

The mercury droplet noise generators produce noise by discharging a charged mercury droplet on contact with a mercury pool acting as a ground plane. The mercury droplet acts as a spherical quarter-wave antenna and radiates energy on discharge. Random or pulse-type noise is generated by controlling the type of flow and size of the droplets. Two basic types of charging the mercury droplets are the induction and the self-charging type. The induction type of charging involves the dropping of small mercury droplets formed by flowing mercury through small parts. The droplet falls through an intense electric field at high velocity, becomes charged, and is discharged on contact with the ground of mercury. In the self-charging type, mercury droplets are charged as a result of the impact of a high-velocity droplet on a piece of dielectric. On impact, the drop flattens, gathers charge, regains its normal spherical shape due to surface tension and rolls off the dielectric fully charged, and on contact with the ground plane is discharged. Frequency range in coaxial units is in order of audio to 5,000 megacycles per second, and for waveguide-type units exceeding 50,000 megacycles per second are achieved.

The Transmission-Line Method of Testing Loop Receivers; *C. E. Kilgour (Crosley Division, Avco Manufacturing Corporation, Cincinnati, Ohio).*

To meet the need for a method of measuring over-all sensitivity of radio receivers having large loop antennas, especially standard broadcast receivers mounted in console cabinets where the loop is as large as the cabinet back, the standards committee of the Institute of Radio Engineers has recently adopted "The Transmission Line Method." Quoting the standard: "The method consists of stretching a . . . wire . . . between two insulators at a uniform distance below the roof (the inner shield) of the screen room and parallel to the longest dimension of the room." One end of the line is connected to a signal generator and the other to the screen of the room through a resistor equal to its characteristic impedance.

Assuming a room with dimensions a very small part of a wave length and with walls which are perfect reflectors, the end walls serve to extend the length of the line so that the field in the absence of the floor, ceiling and side walls has cylindrical co-ordinates and is inversely proportional to the distance from the line. A consideration of all the possible reflections from these last four surfaces yields a summation of a double infinity of terms. If the width of the booth is greater than the distance from line to floor, all terms due to side walls may be neglected leaving only those due to ceiling and floor. The first eight of this set follow:

$$E=60\,I\left[\frac{1}{x}-\frac{1}{2c+x}+\frac{1}{2f-x}-\frac{1}{2f-x+2c}+\frac{1}{2f+x+2c}-\frac{1}{2f+x+4c}+\frac{1}{4f-x+2c}-\frac{1}{4f-x+4c}+\cdots\right]$$

E is the field strength in volts per meter, I the current in the line in amperes, f the distance from line to floor, c the distance from line to ceiling, and x the distance to any point directly below the line. The terms may be combined:

$$E=60\,I\left[\frac{2c}{x(2c+x)}+\frac{2c}{(2f-x)(2f-x+2c)}+\frac{2c}{(2f+x+2c)(2f+x+4c)}+\frac{2c}{(4f-x+2c)(4f-x+4c)}+\cdots\right]$$

At floor level $x=f$

$$E=60\,I\left[\frac{4c}{f(f+2c)}+\frac{4c}{(3f+2c)(3f+4c)}+\cdots\right]\cong 240\,\mathrm{I}\,\frac{c}{f(f+2c)}\;1+\frac{1}{9}+\frac{1}{25}+\frac{1}{49}+\cdots$$

This last has been extended to the equivalent of 16 terms of the original series. The use of a large number of terms is probably not justified because of phase shift and reflection loss.

The committee recommends the use of four terms of the original series, with integration of the field over a large loop to obtain the average field strength.

The committee also recommends that the line be terminated in its characteristic impedance. However, it would appear that since the line is a very small part of a wave length, standing waves are no problem and it is sufficient that the loading resistor be large enough to swamp out the inductance of the line. It is necessary only that the current through the line be calculated readily from the voltage reading of the signal generator. The loading resistor may be placed between the generator and the line with the far end of the line connected directly to the screen, thus reducing the unwanted field due to line potential. With some signal generators, it is possible to place the final shunt resistor of the output attenuator in series with the transmission line, and tune out its inductance with a series capacitor thus securing maximum field strength.