Measurement of Near Fields of Antennas and Scatterers

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Abstract—**A tutorial review of techniques used for the measurement of near-fields of antennas and scatterers, and an extensive bibliography of the readily available literature in this area, are presented.**

I. Introduction

MOST of the experimental measurements made on antennas are concerned with the determination of the fundamental characteristics of the antenna that have to do with their immediate application: the input impedance, the far-field radiation pattern, and the gain or the efficiency. There are however, occasions when it is desirable to have in addition to, or in lieu of, the preceding properties, information about the current or charge distribution on the antenna, and/or the distribution of the near fields, i.e., the fields in the immediate vicinity of the antenna.

The near fields of the antenna can be conveniently divided into the "reactive near field" and the "radiating near field" [1]. The former may be considered to be that region immediately surrounding the antenna, with a usual outer limit approximately a wavelength or less. The latter is that region, beyond the reactive near field, in which the radiation pattern is dependent upon the distance from the antenna. This latter region is usually considered to extend to a distance $2D^2/\lambda$ from a uniformly illuminated radiating aperture of diameter D, and much greater distances for apertures excited with noncophasal distributions.

These two regions may present different measurement problems. Measurement of the current or surface charge distributions on the surface of an antenna involves measurement of the normal electric field and tangential magnetic field very near this surface. Similarly, the measurement of aperture field distributions usually are made within or very near this aperture. These measurements are thus usually made in the reactive near field and the requirement that they be made in very close proximity to the antenna may dictate the use of very small probe antennas to sample these fields.

There is a large class of measurements which can be made at somewhat greater distance from the antenna but still well within the radiating near field. This would include, for example, radiation measurements made on very large apertures or arrays where it is not possible or practical to make measurements at a distance great enough to be in the far field. We will consider, in order, measurements made in these two regions, with primary emphasis upon the reactive near field region.

From an experimental point of view, the problem is to design or obtain probes or receiving antennas with associated apparatus for determining the amplitude, phase, and polarization of the electromagnetic field at any point in space in the immediate vicinity, including the surface (or surfaces), of the radiating antenna. The techniques to be discussed are equally suitable for measuring the fields from scattering bodies.

The availability of the excellent phase and amplitude receivers, vector voltmeters, and network analyzers, which permit direct measurement and recording of the relative amplitude and phase of an RF signal, has eased or eliminated many problems for the experimenter. Therefore we will limit the discussion on instrumentation to a few systems that are of particular interest.

II. Measurement of Surface Current and Charge Distribution

A. Sampling Probe

At the boundary between a medium of high conductivity and free space, the electromagnetic field equations reduce to simple form. On a perfectly conducting surface, the surface-charge density σ, and the surface-current density J are related to the electric ($\bar{E}$) and magnetic ($\bar{H}$) fields by the boundary conditions

$$n\cdot\bar{E} = \sigma/\epsilon_0 \qquad (1)$$

$$n \times \bar{H} = \bar{J}$$

where n is the unit normal directed outward from the surface and ϵ_0 is the permittivity of free space.

In the electrostatic case, and for periodically time-varying fields in the usable range of radio frequencies, the normal component of the electric field may be considered to be zero within all high conductivity materials. Thus the surface charge density may be measured by a probe whose output is proportional to the magnitude of the normal electric field at the medium boundary. Similarly, the depth of penetration of the tangential magnetic field in a good conductor is so small in and above the VHF range of frequencies that the true current

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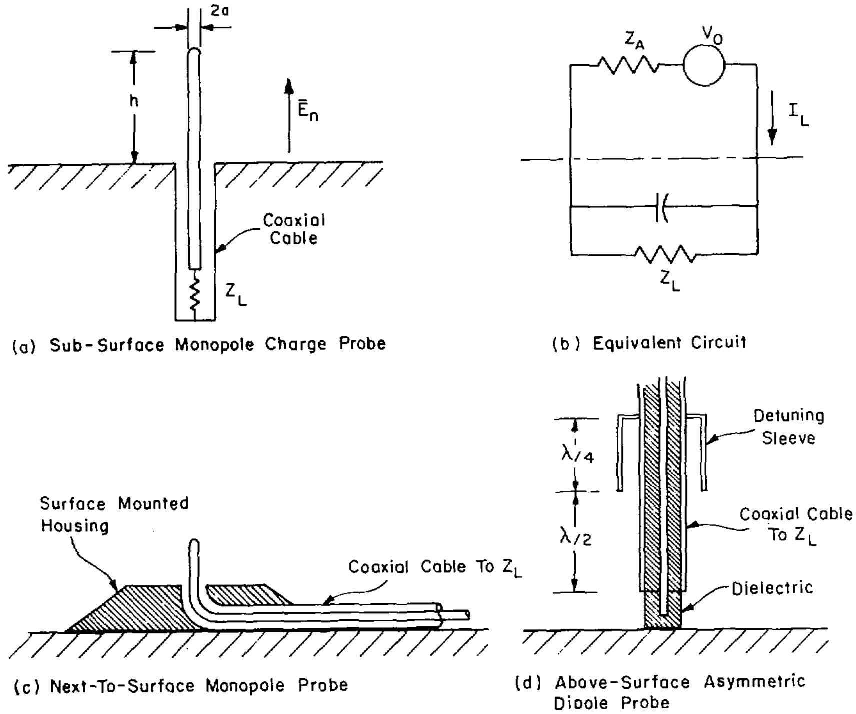

Fig. 1. Charge probes.

distribution is very closely approximated by the surface distribution on a perfect conductor. Thus if the output of a probe is proportional only to the magnitude of the tangential magnetic field it may be considered to be a measurement of the true surface current.

Ideally it is desired to measure the unperturbed normal electric and tangential magnetic fields on arbitrarily shaped surfaces. Suitable probes for the measurement of these quantities would be an infinitesimal electric dipole and an infinitesimal magnetic dipole. The small finite size electric dipole is a good physical approximation of the former; the small finite size loop antenna may, under restricted conditions, provide a physical approximation of the latter.

For accurate measurement of the field intensity distribution near a radio-frequency source, the probe must meet the following requirements [2].

1) Any distortion of the fields by the probe and associated equipment must not seriously affect the accuracy of measurement.

2) The aperture of the probe must be small enough to essentially measure the field at a point.

3) The probe must have the desired polarization to a high degree of accuracy. Since it is usually desired to measure some linearly polarized component of the field, it would be desirable to have the probe linearly polarized.

4) The probe must deliver a signal voltage large enough to permit accurate measurement.

It is axiomatic, however, that if the probe extracts energy from the field for measurement purposes the field is no longer unperturbed. The problem is to bring about a compromise between the maximum allowable perturbation of the fields and the extraction of the minimum energy that will permit the desired identification of the characteristics of the fields.

The measurement of the current and charge on linear antennas dates back at least to 1935 when Gihring and Brown measured the current on models of broadcast towers [3]. Early measurements were also made on antennas consisting of essentially linear elements by other investigators. In the period following the World War II, however, methods of probing the near fields of transmission lines, antennas, and aircraft models were the subject of a series of extensive investigations by R. W. P. King and his associates at Harvard University, and much of our present knowledge of the characteristics of near-field probes is due to the work of this group.

B. *Monopole—Charge Probe*

The normal component of the electrical field on a metal surface can be measured by a short, cylindrical wire receiving antenna oriented parallel to this field. If the geometry of the metal surface permits it, this may consist of a simple extension of the center conductor of a coaxial cable protruding through a hole in the metal surface as shown in Fig. 1(a).

This center conductor is a linear antenna loaded with a transmission line terminated in an impedance Z_L. The electric field normal to the metal interface E_n will excite a current in the load. This antenna may be analyzed in terms of the simplified equivalent circuit of Fig. 1(b) where the antenna has been replaced by the ideal Thevenin generator with open-circuit voltage V_0 in series with Z_A the input impedance of the antenna when driven.

The use of a subsurface probe protruding from the underside of a metal surface, to measure the spatial

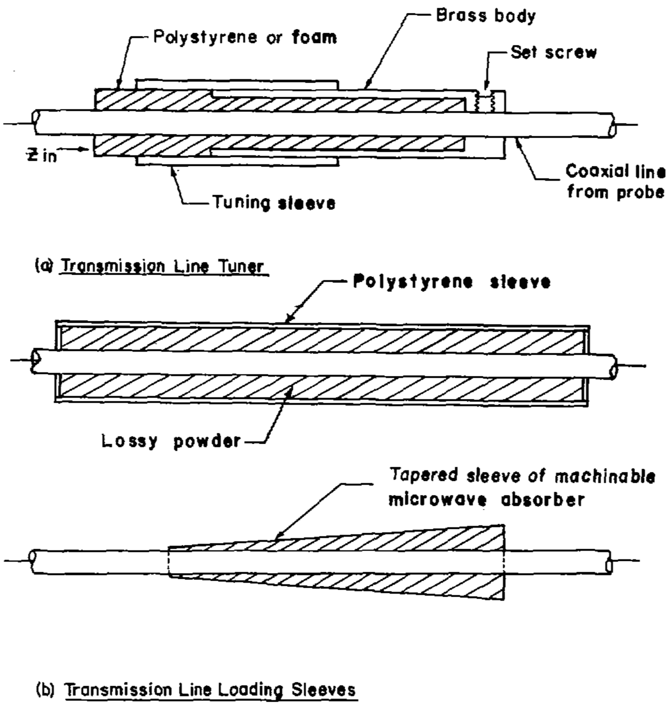

Fig. 2. Chokes and loading sleeves to suppress current induced on exterior surface of coaxial line.

distribution of charge density, is only possible if the surface may be perforated or slotted to allow movement of the probe. If the presence of the holes or if the slot is objectionable, these can be covered when not occupied by the probe. However, the utility of this image plane system is limited and it frequently becomes necessary to use a probe system which is external to the surface.

For flat metal surfaces this may take the form of a probe protruding from a small surface mounted housing with bevelled edges as shown in Figure 1c. Reynolds [4] termed these "next-to-surface" probes and used short monopoles, loops and cavity backed slots mounted in this form of housing. Although there is some distortion of the magnetic field due to the change in the contour of the surface caused by the housing and connecting coaxial cable, such techniques have been used at VHF and UHF frequencies.

A more flexible charge probe can be constructed in the form of an asymmetrical dipole as shown in Fig. 1(d). A rigid coaxial line is positioned perpendicular to the surface and the protruding center conductor, with the outer conductor, forms a receiving antenna. Because there are currents induced on the outer surface of the coaxial line, and there is direct capacitive coupling to the lower rim of this outer surface, the equivalent circuit is complicated. This circuit would incorporate three generators, and by superposition the probe output would be the sum of the complex voltages impressed upon the load Z_L by these generators [5]. Since only coupling to the center conductor is desired, the capacitance between the lower rim of the line and the surface should be kept low. Coupling to the outer surface may be minimized by placing a quarter wave detuning sleeve, such as shown in Fig. 1(c), a half-wavelength from the end of the coaxial line.

Since the probe should couple to the normal component of the electric field, it is necessary that it be maintained perpendicular to the surface and a constant distance from the surface as it is moved. The latter can be insured by extending the coaxial line insulation past the end of the center conductor for a short distance and then maintaining contact between this insulation and the surface to be probed. Precise positioning and movement of the probe requires a carriage and track mechanism driven by a lead screw or an equivalent mechanical arrangement. Since this mechanism is in the field of the antenna its backscattering cross section must be kept small. Most parts of the probe positioning mechanism can be made of polystyrene, foam, wood, or other low dielectric constant material, and whenever possible this mechanism should be shielded with RF absorbing material.

It is essential to eliminate resonant lengths of metal from the system. The radial arm on which the probe is mounted is usually a small diameter metal tube since the metal transmission lies within or on it anyway. Induced currents on this metal conductor radiate a field which distorts the original field distribution in the region of the probe. This effect can be reduced by placing quarter-wavelength sleeves or chokes at intervals to prevent these induced currents from becoming resonant and building up to large values. The sleeves must be adjusted to a resonant length, which is slightly less than quarter-wavelength, and be of a diameter appreciably larger than that of the outer conductor of the coaxial line to be effective. If it is desirable to change the frequency of measurement easily, the currents on the outside of the probe line can be dissipated in properly placed loading sleeves of lossy material such as shown in Fig. 2. Coating the line with carbon based lacquer has sometimes been helpful.

C. Loop—Current Probe

1) Simple Loop: The most common device used for the measurement of high-frequency magnetic fields and surface currents is the loop antenna, and the use of the small loop as a probe has been treated in detail in [5]–[8].

In its simple form the loop antenna consists of a single turn of wire as shown in Fig. 3(a). For the moment, we assume that this loop is immersed in a plane linearly polarized electromagnetic wave in free space, and that the loop is oriented such that it lies in the yz plane, perpendicular to the magnetic field B_x which is considered to be constant over the area of the loop. The loop circumference is assumed small in wavelengths, and the load impedance can be considered to be connected across a slice or gap in the contour of the loop. Under these conditions the current I_H induced in the loop by the magnetic field, is equal in magnitude at all points on the contour. The voltage V^e across the gap may be obtained by equating the line integral of the electric field around

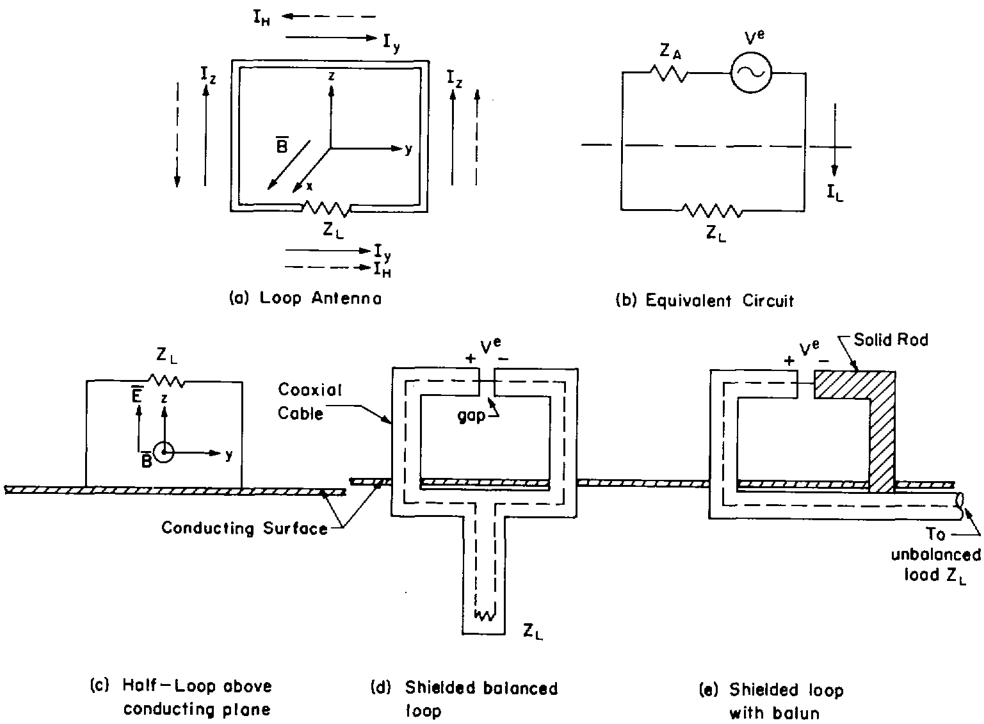

Fig. 3. Loop antenna probe for measurement of surface current.

the closed contour, to the time derivative of the magnetic flux through the loop.

For the calculation of the current I_L through the load, the equivalent circuit of Fig. 3(b) may be used. The loop is replaced by an equivalent generator with open circuit voltage V^e in series with its internal impedance equal to the self-impedance of the loop at the input terminals Z_A and a load impedance Z_L. In addition to the circulating or loop currents due to the magnetic field there will be codirectional, or antenna, currents induced in the sides of the loop that are parallel to the electric field. If the electric field is linearly polarized and oriented along the z axis, these currents will not contribute to V^e. If the electric field, or a linearly polarized component of this field, is oriented along the y axis there will be a current I_y excited in the loop due to the asymmetry about the xy plane caused by the presence of the load. The total current through the load will be

$$I_L = I_H + I_z + I_y. \tag{2}$$

Since we can construct a loop with the load located on one of the lines of symmetry, one of these components (I_z in Fig. 3(a)) will vanish at this line and can be ignored. If then the electric field to be measured is linearly polarized in the plane of the loop, the loop may be oriented so that I_y is zero and I is equal to I_H. If, however, the field is elliptically polarized, as it is in many near-field measurements, no such orientation is possible unless the loop is constructed with a second plane of symmetry. This requires symmetrically placed loads on opposite sides of the loop.

2) Shielded Singly Loaded Loop: For surface current measurements on flat metal surfaces, a loop such as in Fig. 3(c) will provide the symmetry required to minimize the response to the electric field. This may be made into a practical probe antenna as in Fig. 3(d) by constructing the loop from coaxial cable with a narrow gap in the outer conductor or shield. The advantage of this shielded construction results from the fact that the current excited on the loop by the external field flows on the symmetrically constructed outer surface of the shield. Energy is transferred to the load Z_L as a result of the field set up across the gap. When the gap is small it approaches the condition of the idealized slice load impedance of Fig. 3(a) and because of symmetry I_L will be due only to I_H.

The shielded loop may be converted for use with an unbalanced coaxial line by incorporating a simple balanced-to-unbalanced (balun) transformer as a part of the loop structure. The center conductor of the coaxial line is terminated on the outer conductor, or shield, of the opposite half-loop at the gap. Energy will still be transferred to load Z_L in the desired mode since the configuration at the gap merely represents a series impedance to the loop currents. Somewhat greater symmetry can be achieved at the gap by letting the center conductor extend into the opposite half-loop and making this half loop a coaxial line that is open or short circuited. It is possible to use this line extension as a tuning stub in some cases, but this is frequency sensitive and not usually convenient if the loop is very small in wavelengths.

As with the subsurface charge probes, the utility of these probes is limited to those cases where the position of the probe is fixed, or if it must be moved, to those cases where a slot can be cut in the conducting surface. On large flat surfaces a next-to-surface mounting may be practical. Where neither of these are practical the ex-

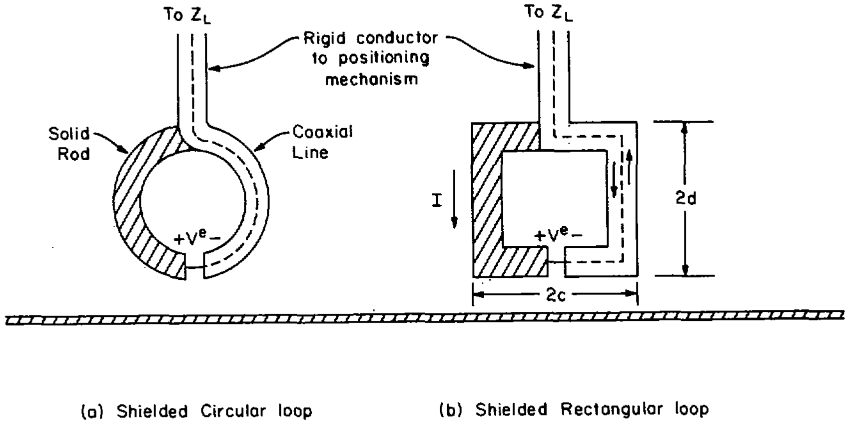

Fig. 4. Above-surface shielded single-loaded loop probes.

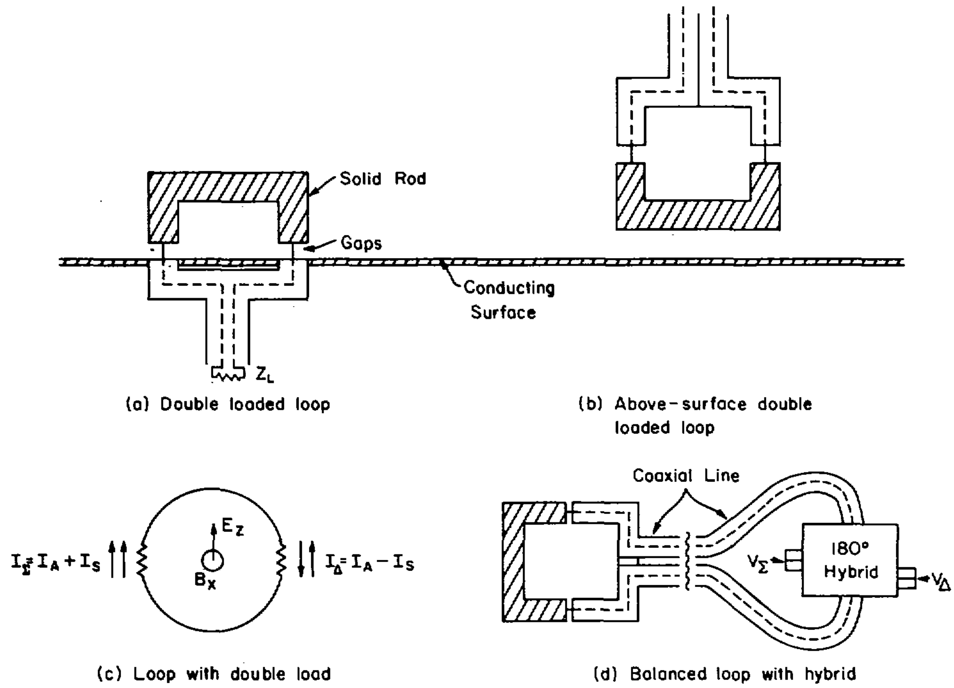

Fig. 5. Double-loaded shielded loop probes.

ternally supported probes of Fig. 4 may be used. The comments made in regard to the positioning mechanism and use of detuning sleeves for the externally supported probes again apply.

The magnetic sensitivity of small loops is proportional to the loop area and hence the loop must be of sufficient size to provide a useful output. Balanced against this is the fact that the output of the probe is proportional to the average value of the magnetic field through the loop aperture, hence the aperture of the probe must be small enough to essentially measure the field at a point. The square loop has been shown to be the optimum shape for minimizing the averaging error for a given magnetic sensitivity in a general incident field [5]. However, at UHF and microwave frequencies, where the shielded loops must be formed of coaxial lines of small cross section, the circular loop is usually the easiest to construct. While the area of a circular loop of diameter $2d$ is approximately 20 percent less than that of a square loop having sides $2d$, so that its sensitivity is not optimum, this disadvantage is small. It is compensated by the fact that the undesired response of the circular loop to the electric field is also less because the perimeter of the circular loop is reduced in the same ratio as its area [7].

Considerable care must be exercised in the use of the singly loaded loop, since the electric field might not be normal to a plane surface at a very small distance from the metal surface at edge discontinuities such as apertures in the surface. In addition, on even the relatively simple linear antenna, although the Ampere–Maxwell equation indicates that the tangential magnetic field at the surface of the antenna is proportional to the total axial current on the antenna, the presence of a non-vanishing longitudinal electric field at points away from the antenna surface means that the magnetic field away from the surface is no longer exactly proportional to the current.

Most probes measure the fields a finite distance, however small, away from the surface.

3) Doubly Loaded Loop: Symmetric double loaded loops are shown in Fig. 5. These loops may be round or rectangular. The total current in Z_L due to the fields across the two gaps can be resolved into symmetric and antisymmetric components. As indicated in Fig. 5(c) the symmetric or in-phase loop currents (I_S) are proportional to the magnetic field B_x, and the antisymmetric or out-of-phase loop currents (I_A) to the electric field E_z. Therefore, the sum of these components I_B is proportional to B_x and the difference of these currents I_E is proportional to E_z. In a practical system I_B and I_E can be obtained from a 180° hybrid as in Fig. 5(d). However, since the hybrid junction will not have perfect isolation between the sum and difference ports, the available currents will be

$$I_B = I_\Sigma + \gamma I_\Delta \quad (3)$$

$$I_E = I_\Delta + \gamma I_\Sigma \quad (4)$$

where γ is the isolation of the hybrid.

4) Loop-Probe Error Ratio: To compare the ability of different probes and the connecting circuit to discriminate against electric field effects, Whiteside and King [8] defined a "system error ratio" e_n. This is the ratio of the output current due to a unit electric field parallel to the loop to the output current due to a unit magnetic field through the loop. The subscript n indicates the number of symmetrical loads on the loop. The error ratios are dimensionless constants given by

$$e_1 = K_{E_1}/K_{B_1} \quad (5)$$

$$e_2 = K_{E_2}/K_{B_2} \quad (6)$$

where K_E and K_B are sensitivity constants of the probe to the electric and magnetic fields and are dependent upon the probe geometry alone.

Whiteside [7] determined that for the singly loaded loop, the error ratio e_1 is independent of load impedance and, for wire radii less than 0.003 wavelengths, independent of wire size. For the doubly loaded loop, e_2 is a function of load impedance and wire size, with small wire and small loads reducing the value of e_2. The error ratio e_1 of the singly loaded loop decreases markedly with loop diameter. The measurement error in a system using a doubly loaded loop is much less than e_1 by an amount dependent upon the isolation through the hybrid used as a combiner. Measured values of the magnitude and phase of the system-error ratio e_n in dB (i.e., $20 \log e_n$) are plotted in Fig. 6 for square and circular loops as a function of the side length or diameter in wavelengths. These curves are based upon γ of −26 db. Commercial coaxial hybrids are readily available for use over octave bandwidths up to 2400 MHz with γ greater than 45 dB and up to 10 500 MHz with γ greater than 35 dB. In the HF, VHF region hybrids are available for use over several octaves with γ greater than 30 dB. In waveguide, matched magic tees are available with 40 dB of isolation.

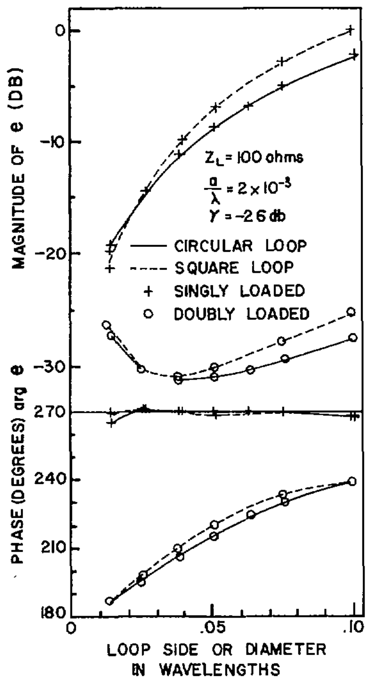

Fig. 6. System error ratio e, ratio of output current due to unit electric field parallel to loop, to output current due to unit magnetic field through loop. (a = wire radius) (after Whiteside and King [8]).

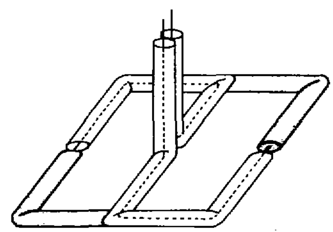

Fig. 7. Symmetrical bridged loop probe.

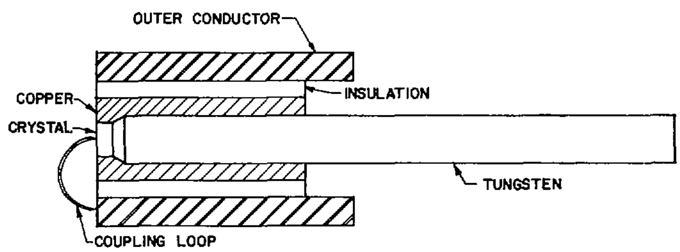

Fig. 8. Combined magnetic loop-diode detector (after Degenford *et al.* [10]).

The values of error ratio shown in Fig. 6 were measured with image plane loops such as in Fig. 5(a). The presence of the transmission line leading to the above-surface loops will increase this ratio. For a singly loaded loop the load gap should be directly opposite the transmission line. With a doubly loaded loop this is not possible unless the loop is constructed with a conducting bridge as shown in Fig. 7. The bridged loop has a reduced sensitivity to the electric field and hence the error ratio is reduced a few dB [6]. The improvement over the unbridged double loaded loop is less than 1 dB for large probes or for image probes, but may reach 5 dB for small free-space or above-surface probes. This reduction in e_2 may be of importance in situations where the cross coupling of modes is significant.

5) Loop for Use at Millimeter Wavelengths: A combined magnetic loop, diode detector has been devised for use at millimeter wavelengths [10]. As shown in Fig. 8, a crystal chip is soldered to the end of a tungsten rod which becomes the center conductor of a coaxial line. A small diameter tungsten wire formed into a small loop between the outer conductor and the crystal serves as a coupling loop and whisker for a point contact diode, An experimental unit has been tested at 75 GHz. It should be possible to construct units to go much higher in frequency. This is an unshielded loop and hence must be used with some care to minimize the response to the electric field. Its disadvantages in this respect may be outweighed by its efficiency and increased sensitivity over other types of probes and waveguide mounted diode detectors at these extremely short wavelengths.

6) Current and Charge Probes Using Scattering Techniques: It should be mentioned that it is possible to measure the current and charge on metal surfaces and antennas by using a radically different technique. This technique depends upon using the properties of the probe as a scatterer. A discussion of the use of this type of probe to measure current and charge distributions will be delayed until the technique is developed in the next few sections.

III. Measurement of Aperture Distribution and Near Fields

Aperture field distributions can be measured by two general classes of probes. The first class provides a direct sample of the field. For example, the magnetic field may be measured by the above-surface loop probes considered in a preceeding section and the electric field may be measured by an electric dipole or waveguide horn antenna. The second class depends upon the measurement of the backscattering from a small probe which may or may not have a direct electrical connection to the measurement circuit.

A. Probes for Direct Measurement

1) The Electric Dipole: Consider a linear antenna of length $2h$ and diameter $2a$, center loaded with a transmission line terminated in its characteristic impedance Z_L as shown in Fig. 9(a). For small diameters this dipole is a good electric field probe, responding only to the tangential component of the electric field.

The measurement of near fields is, in general, affected in a complicated way by the presence and characteristics of the probe antenna. This problem, which will be discussed later, has been considered by a number of workers and correction of the measured data is possible. Without such correction Borts and Wooton have shown that the dipole as a probe should certainly not be larger than a half-wavelength [11], [12]. Although the sensitivity of the dipole is a maximum for a resonant length, the resonant antenna scatters much more energy than a shorter dipole, and such a long probe responds to an average field and not a field at a point. For this reason the probe is usually kept small. Similarly, although the thick antenna is somewhat more sensitive, the thin antenna has a greater resolution for the determination of the polarization characteristics of the electric field. If the probe is known to be linearly polarized, the polarization of the electric field may be obtained by rotating the probe about axis A in Fig. 9(a), and recording the amplitude of the field as a function of this rotation, when this axis, and hence the plane of rotation, is oriented for maximum response.

The balanced dipole must be connected to the unbalanced line through a suitable balun. This may take the form shown in Fig. 9(c), where the center conductor is connected to one dipole arm and the outer conductor to the other, with this outer conductor slit for quarter-wavelength [9]. This balun is frequency sensitive, and if it is desirable to use the probe over wide bandwidths, another type must be used.

A technique that may be used over octave and wider bandwidths is indicated in Fig. 9(d). The dipole arms are connected to equal length individual coaxial lines, which in turn are connected to the side ports of 180° coaxial hybrid. The series port and the two side ports of this hybrid function as a shielded balun and since the fourth port (the sum port) is not necessary it should be terminated in a 50-Ω load. The two center conductors of the coaxial lines become a balanced, shielded transmission line of 100-Ω characteristic impedance [13]. The outer conductors of the two coaxial lines should be in contact with each other, or bonded together in the region near the dipole to prevent these outer conductors from acting as a two wire line that can become resonant in length.

The two lines may be enclosed in a small diameter tube of length l to provide support for proper positioning. As for all such supported probes, detuning or absorbing sleeves are usually necessary to minimize the backscatter from this metal tube. The dipole and balun must be carefully constructed and the connecting lines to the hybrid matched in length to obtain satisfactory results.

It is usually advisable to check the polarization discrimination of a dipole probe by rotating it in a known linearly polarized field. The fields inside an open-ended reduced height rectangular waveguide excited in the dominant mode provide a convenient check on the polarization of probes which are small compared to the dimensions of the waveguide.

2) Wireless Probes: The conventional probing arrangements that have been discussed present a number of difficulties. The probe, and particularly its associated metal clad coaxial cable or waveguide, are a major source of field perturbation. Methods of minimizing this distortion have been discussed, but there is always a question of the reliability of the measured data.

At HF and VHF frequencies, it is possible to overcome many of these difficulties by telemetering the data from

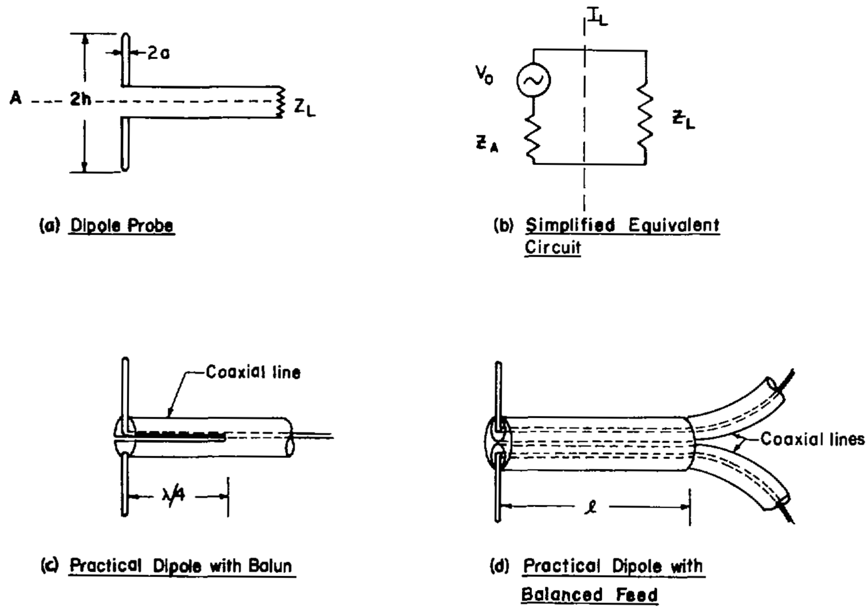

Fig. 9. Dipole electric field probes.

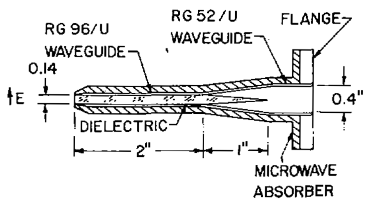

Fig. 10. Waveguide probe (after Richmond and Tice [2]).

the probe. Iizuka [14] developed a small battery operated tunnel diode oscillator to be mounted on a dipole probe for measuring the radiation from a test antenna at 114 MHz. In this system the radiated signal from the test antenna, amplitude modulated at 1000 Hz, was received by the probe where it was detected. The 1000 Hz signal was used to frequency modulate a 7 MHz tunnel diode oscillator. The 7-MHz FM signal was fed to the probe dipole through a duplexer consisting of two-tank circuits, radiated, and received by a remote receiving antenna. The probe was designed for use under water as well as in free space. With the dipole replaced by a small shielded loop, the probe would telemeter either the H field or the current distribution along an antenna.

3) *Waveguid Probe:* In the microwave and millimeter region, small horns or open-ended waveguides are used as probes. One x-band open-end waveguide probe developed by Richmond and Tice [2] is shown in Fig. 10. The open-end guide was loaded to permit a reduction in cross section, and covered with absorber to reduce reradiation and distortion of the fields.

B. Indirect Methods of Measurement

1) *The Use of Small Scatterer as Probe:* The presence of the waveguide or coaxial lines leading to the probe causes difficulties other than those due to field perturbations. When phase measurements are undertaken, it is necessary to bring the RF signal from the probe back to the receiving apparatus by means of a relatively long waveguide or transmission line. When the probe is moved to obtain the spatial distribution of the fields, there will be an unavoidable flexing of the transmission line or motion of required rotary joints. This movement can cause serious errors in the measurement of the phase of the fields.

To circumvent these difficulties, Justice and Rumsey [15] proposed a method of probing an electric field by means of a small scatterer. The principle of the method rests upon the measurement of the reradiated or scattered field from a small probe as it is moved through the field of interest. The observed quantity is the signal V due to this scattered field, received by the antenna under test. It follows from the reciprocity theorem that this voltage V is related to the electric field $\bar{E}$ at the position of the scatterer [16], [17].

To relate these quantities, we assume the scatterer is very small in terms of wavelengths and use a Rayleigh approximation [18]. In the vicinity of this small scatterer, the electromagnetic field distribution can be calculated as though the wavelength were infinite, making available the results of potential theory. Thus, a very small scatterer of any arbitrary shape immersed in a field $(\bar{E},\bar{H})$, can be characterized as an elementary electric dipole with current moment $\bar{p}$, where

$$\bar{p} = \alpha\bar{E}. \tag{7}$$

The polarizability of the scattering body (α) is, in general, a 3×3 matrix which can be evaluated by considering the problem of the scatterer immersed in a static field $\bar{E}$. For a few simple geometries α can be found analytically. In other cases it has been determined experimentally, sometimes using analog techniques, In the practical case, a convenient scatterer is a very short straight cylinder.

In this case the matrix α reduces to a single term, and $\bar{p}$ is equal to α times E_l, the component of $\bar{E}$ parallel to cylinder axis

$$\bar{p} = \alpha E_l. \tag{8}$$

Thus $\bar{p}$ is proportional to the undisturbed field $\bar{E}$ radiated by the antenna under test. It is assumed that $(\bar{E},\bar{H})$ is generated by impressing a current i_0 across a pair of terminals connected to the test antenna. By reciprocity, there will be a voltage V developed across the terminals of this antenna when it is immersed in the radiated field due to p (i.e., the scatterer) and

$$i_0 V = \bar{p}\cdot\bar{E}. \tag{9}$$

From (8) and (9), we see that

$$V = AE^2 \tag{10}$$

where A is a constant. In addition to the assumption that the scatterer is small, we have assumed that the scatterer stays in the same medium.

Equation (10) indicates that the reradiated field or echo signal V from the scatterer is proportional to the square of the component of the electric field E_l oriented along the scatterer and the phase angle of V is twice the relative phase angle of E_l plus a constant (the phase of A).

Equation (10) assumes the scatterer is sufficiently far removed from all conducting bodies to be essentially in free space. Justice and Rumsey [15] estimated the magnitude of error in the determination of $\bar{E}$ when the scatterer approaches an infinitely large, perfectly conducting sheet. They calculated that this error did not exceed 1 percent, provided the distance from the reflecting surface is greater than the length of the scatterer. In a closely related study, Plonsey [19] took into account the radius of a cylindrical scatterer when investigating the interaction between an active scatterer and an infinite reflecting plane, and obtained the error that could be expected when the scatterer is in close proximity to a reflector. Fig. 11, which has been extracted from Plonsey's work, indicated that we might expect an error of approximately 2 percent for a very small diameter probe of length $\lambda/16$, spaced $\lambda/20$ from a reflector. This error rises to approximately 6 percent for the same diameter scatterer of length $\lambda/8$, and for any length the error rises as the radius of the scatterer increases.

This method of sampling a electromagnetic field has major significance. Since the scatterer can be supported and positioned by very thin nylon thread, it eliminates the metal clad coaxial cable or waveguide leading to the probe, and eliminates phase errors due to flexing of the coaxial cable or those due to movement of waveguide rotary joints in the lines leading from the probe. Given sufficient sensitivity in the receiving system, it permits very small simple probes and hence it is very useful at microwave frequencies. It also permits the measurement of fields in confined space such as inside waveguides, in liquid, and since the scatterer can be slid through very small holes, in solid dielectrics.

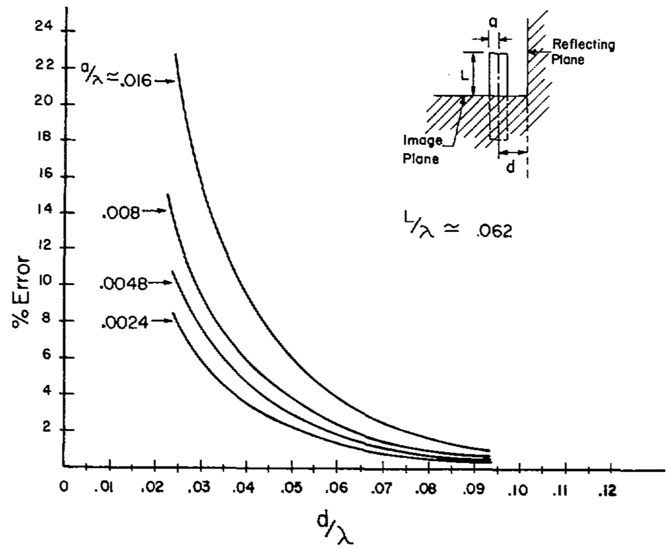

Fig. 11. Error due to interaction of probe of length L with spacing d to infinite conducting plane for various values of probe radius (after Plonsey [19]).

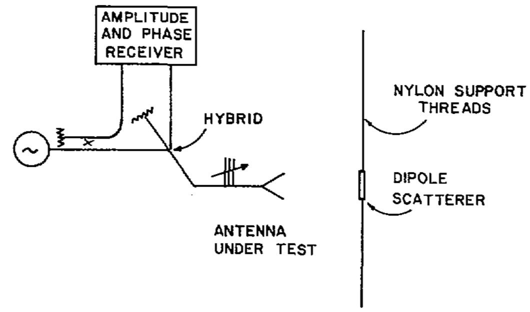

Fig. 12. Instrumentation for measurement of field distribution with small scatterer. If receiver is not available, system described in [15] may be used.

The scattering method of measuring electric-field distributions can be instrumented as shown in Fig. 12. A hybrid junction is used to isolate the receiver or detector from the signal source. The collinear (or side) arms are connected to the antenna under test and to a resistive load, and the E and H arms (series and parallel arms) are connected to the signal source and the receiver. If the reflection coefficient of the load is equal to that of the antenna no power would be transmitted from the signal source to the receiver through an ideal hybrid. In practice, because the signal to be received is very small, the antenna must be tuned until the isolation between signal source and receiver is about 90–100 dB. An undesired signal to the receiver due to imperfect balance of the hybrid junction which is 20 dB below the desired scattered signal would still contribute a ± 5 percent error to electric-field measurements [20]. In addition, care must be taken to prevent motion of any bodies, except the scatterer, in the antenna field during measurements.

To overcome the disadvantages of this system, Richmond [20] proposed that the scattered wave be modulated by electronic means. Independently, Cullen and Parr [21] proposed that an electromagnetic field could be measured by perturbing it with a small scatterer and that this scatterer be a spinning dipole. This in effect provides a scattered signal that is modulated by mechanical means.

2) *Modulated Scattering Technique:* If a nonlinear impedance such as a diode is placed at the center of a wire scatterer, and a 1000 Hz square-wave modulation current is applied to this diode through slightly conducting thread leads, as shown in Fig. 13, it can be shown [22] that the observable voltage V is given by

$$V = V_0 - \frac{CE^2l^2}{Z + Z_L} \tag{11}$$

where the first term is a continuous signal which will not be detected by the system to be used, C is a constant proportional to the source excitation, l and Z are the effective length and input impedance, respectively, of this dipole scatterer and Z_L is the diode junction impedance. This impedance Z_L, the load impedance at the effective terminals of the scatterer, will change at the modulating frequency and, hence, the resulting scattered field will be modulated. The magnitude of V is still proportional to the square of the magnitude of the component of $\bar{E}$ along the direction of the diode and the phase of V is equal to twice the phase angle of $\bar{E}$ plus a constant phase shift. This phase shift depends upon Z and the two values of Z_L corresponding to the two states of the square wave modulating signal.

Instrumentation that can be used to monitor this scattered field is shown in Fig. 14. This will be recognized as a coherent detection system with balanced mixer [23], and the voltage output from the selective amplifier at the angular modulation frequency ω_m can be shown to be of the form

$$V \simeq K[E_1E_2^2 \cos (2\phi - \theta)] \cos \omega_m t \tag{12}$$

K is a constant, E_1, and θ the amplitude and relative phase of the signal in the reference arm, E_2 and ϕ the amplitude and relative phase of the signal received from the scatter. If the amplitude of the reference signal E_1 is held constant and its phase adjusted to make $2\phi = \theta$ then $V \sim E_2^2$. The output of the mixer is independent of the phase or amplitude of any unmodulated unbalanced signal.

Other methods of detecting the modulated scattered signal have been proposed [25], however because these systems depend upon differentiation between the modulated scattered signal and unmodulated signals, most of the phase-amplitude measuring sets cannot be directly used.

The modulated scattering technique not only dispenses with the necessity of a transmission line leading from the dipole probe but also relaxes the requirements on tuning and frequency stability. Signals scattered back from supporting frames and other stationary structures are not modulated, therefore, they do not unduly affect the accuracy of the experimental results.

The effects of the nylon supporting threads have been found to be negligible. Similarly, the cotton thread coated with colloidal solutions of graphite, which is used to carry the modulation voltage, has negligible effect on the measurements if the resistance near the dipole is approximately 200 000 Ω/ft or greater. Three inches of coated cotton thread at both ends of the dipole should provide sufficient isolation from the rest of the audio circuit [20] at microwave frequencies.

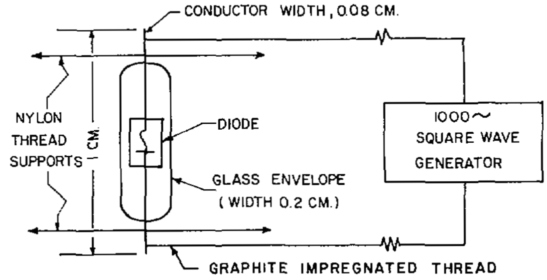

Fig. 13. Diode used as modulated scatter for field measurements (after Richmond [20]).

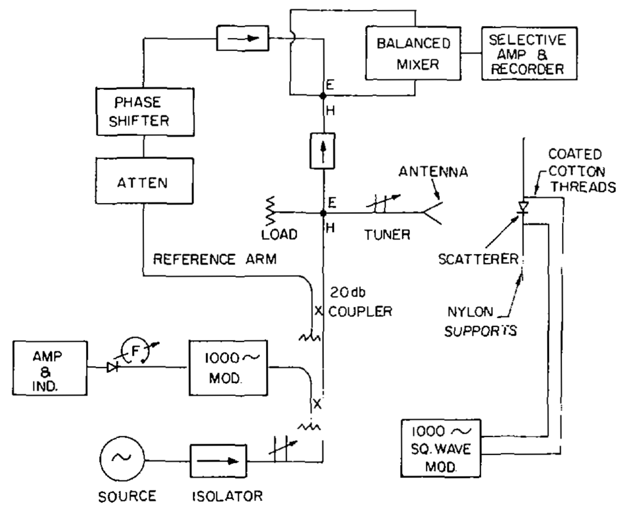

Fig. 14. Instrumentation for measurement of field distribution with small modulated scatterer.

The backscattered field from the scattering probe can be enhanced by making the probes resonant or by using active elements to load the probe [26]–[28].

It is possible to measure the magnetic field by using a scattering loop formed of two diodes [22], [24]. As far as the magnetic field is concerned, the diode loop can be considered simply as an ordinary conducting loop loaded with two impedances due to the diode junctions. This is an unshielded loop which will respond to both the $\bar{E}$ and $\bar{H}$ fields. It is possible, however, to use a single diode scatterer to determine the direction of $\bar{E}$, and with the direction of $\bar{E}$ known, the $\bar{H}$ field can be determined by properly orienting the diode loop probe. As long as the direction of the diodes are kept perpendicular to $\bar{E}$, the $\bar{E}$ field effect is minimized or not detected.

If the presence of some RF circuitry at the probe can be tolerated, King [29] has proposed a scheme (with an excellent discussion of the possible errors involved) in which the signal from the probe is single sideband

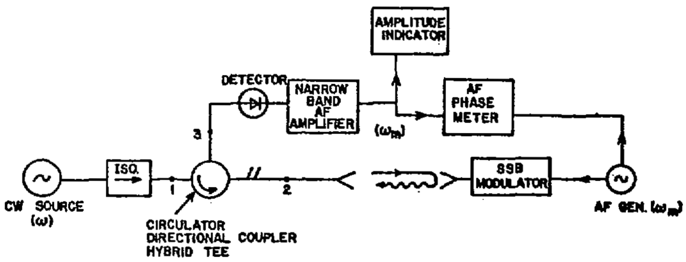

Fig. 15. Instrumentation for real-time measurement of field distribution with SSBSC homodyne system (after King [25]).

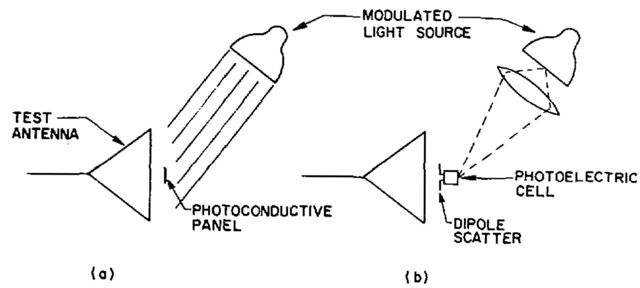

Fig. 16. Light modulated scattering probes for measurement of electric fields. (a) After Potts [30]. (b) After Vural and Cheng [34].

modulated (SSBSC) and then reradiated as shown in Fig. 15. Under these conditions the output of the detector is proportional to

$$V \simeq KE_1E_2^2 \cos [\omega_m t + 2\phi] \quad (13)$$

and the amplitude and phase of the field is immediately available. The reader is referred to [23] for a general discussion of the effect of different types of modulation in these systems. Using this "SSBSC Homodyne" system with low-noise detectors King has demonstrated sensitivities of -110 to -120 dBm [29].

There is an effective method of modulating the field, which is scattered by the probe, without making any electrical connection to the probe. As shown in Fig. 16, a small photoelectric panel or cell is used in place of the diode scatterer [30]–[34]. Focused pulses of a light beam shine on the cell at an audio rate and the modulating mechanism relies on the periodic change in conductivity of the photoconductive material. Wohlleben and Hennig [33] have studied the static and dynamic "reaction-factor" (defined to be one minus the magnitude of the reflection coefficient) of light modulated photoconductor probes as a function of the maximum dimension of the probe.

3) *Perturbation Technique:* An interesting method of obtaining the scalar amplitude distribution in the aperture of an antenna has been used by Cornbleet [35]. The method is based upon the single precise measurement of the aberration of the direction of the peak of the radiated beam caused by partially obscuring the aperture with a reflectionless dielectric sheet, strip, or disk, of known small phase-insertion delay.

4) *Measurement of Surface Charge and Current Distribution with Modulated Scatterers:* Within the limitations implied by Fig. 11, it is possible to measure the charge and current distribution on large surfaces of antennas or scattering bodies, and the surface impedance of plane stratified media [36]. In the HF, VHF, and UHF range of frequencies, it is practical to measure these aforementioned distributions on the surface of thin linear antennas as well. The scatterers may take forms shown in Fig. 17.

A small shielded coplanar loop, loaded with a photocell, has been used by Iizuka to measure the current distribution on monopole and dipole antennas at 600 MHz [31].

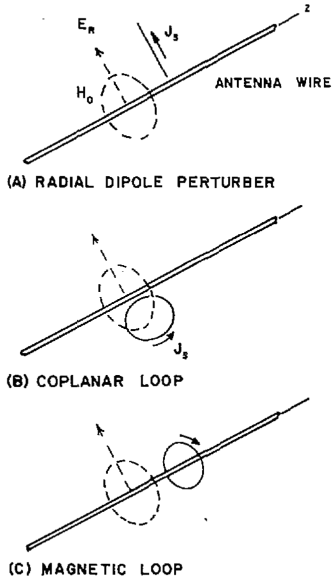

Fig. 17. Perturbers or scatterers that can be used to measure current or charge distribution on thin linear antenna (after Cory and Fenwick [37]).

Cory and Fenwich [37] extended this technique to the HF range by measuring the current distribution on one arm of a loaded V antenna at various frequencies in the 2–30 MHz range. They used a coaxial loop perturber which was effectively opened and shorted by a diode switch. The battery operated perturbing (i.e., scattering) package consisted of a magnetic loop, a diode switch, switch driver circuits, an FM receiver and a small monopole antenna. The switching signal for the diode was obtained by modulating a 100–108 MHz FM generator with a 1615-Hz audio signal. This 108 MHz signal was then transmitted to the perturbing package where it was received by the monopole on this package. At the input to the antenna under test, the transmitted and reflected signals were combined in a coaxial T rather than a hybrid. To discriminate or identify the small scattered signal in the presence of the strong transmitted signal, the combined signals were fed to a spectrum analyzer. Thus, the carrier and its sidebands could be displayed. The amplitude of one of the two sidebands was monitored since the components of the signal at the sideband frequencies were due to the modulated scattered signal.

IV. Measurements in Radiating Near Field

Measurements of the far-field radiation pattern of any aperture involves probing the field at a constant radius far enough away that the aperture approximates a point source. This distance depends upon the degree of approximation desired. A typical criterion is that distance at which a maximum phase error of 22.5° or $\frac{1}{16}$ wavelength occurs over a region of the same size as the aperture plane [38]–[40]. For a uniformly illuminated aperture of diameter D, this distance is given by $R = 2D^2/\lambda$. At this distance the antenna directivity is within 1 percent of the value at an infinite distance. For non-cophasal apertures this $2D^2/\lambda$ criterion can be very unreliable, and much greater distances may be required.

As the observer moves nearer to the antenna under observation two effects are noted. The main beam broadens, the nulls between lobes fill in and the sidelobes are slightly raised [41]. These effects become progressively more pronounced as the distance decreases until a point is reached at which the main beam tends to bifurcate into two beams, depending upon the aperture and aperture distribution [42]–[46].

Since in the near field the angular distribution of the radiated fields, i.e., the radiation pattern of the antenna under test is a function of the distance between this antenna and the receiving or probe antenna, it is highly desirable that pattern measurements be made beyond this near field. For large antennas the required distance to the far field may be thousands of feet and obtaining a suitable test site becomes a formidable problem. Many attempts have been made to overcome this problem. These attempts have in the main taken two forms. In the first, the test range or antenna under test may be physically modified. In the second, the measured near-field patterns or near-field distributions are accepted and computational techniques used to predict the approximate far-field pattern.

A. Modification of Test Range

1) *Collimating Systems—Lenses:* In the microwave range of frequencies it is feasible to use a lens to collimate the radiated energy and attempt to simulate a plane wave [47]–[50]. The difficulty with this technique is the fact that available microwave lenses are relatively small, and consequently, diffraction effects are pronounced and the emergent wave deviates from its plane wave characteristics. Reflections at the lens surface can cause multiple scattering but these can be reduced by the use of reflectionless surface matching layers [51]–[52]. An improved lens arrangement uses two lenses [53]. The second lens allows the phase and intensity distribution over the collimator to be suitably adjusted. Calculations have shown that in order to obtain a plane phase front at some preassigned distance from the collimator, a quadratic phase error must be introduced of such a form as to cause the emergent wave to converge towards a definite point in the image space. A Gaussian amplitude distribution combined with a quadratic phase variation has given a significant improvement over a single lens system [54]. For good results, the collimating lens should in general be several times the size of the antenna under test.

2) *Collimating Systems—Compact Range:* In an ingenious arrangement, Johnson [55] showed that uniform plane waves can be generated across the aperture of a target or antenna by using large paraboloids or parabolic cylinders as collimators. A "compact range," based upon these devices, has been used for antenna and radar reflectivity measurements with good results [56]. This technique would appear to deserve more attention.

3) *Focusing Antenna:* Conventional antennas are focused at infinity. Techniques have been developed for the measurement of far-field patterns in the near field, which depend upon designing the antenna so that it focuses at a finite point in space. These designs produce, in a region about the focal point, the pattern normally found in the far field. The pattern exhibits a depth-of-field behavior analogous to optical lenses. This technique has proved to be practical for large arrays and apertures which can be molded or curved about a spherical surface [57], [58], and for large paraboloidal reflectors [59]–[62].

B. Techniques for Determining Far-Field from Near-Field Data

The extrapolation from measured near-field data to far-field information can be approached from various points of view. In principle, it is possible to measure the field distribution in the aperture of the antenna, and use the Fourier transform relation between the aperture distribution and the radiation pattern [63]. However, the actual field distribution obtained by measurement can seldom be expressed by a simple closed form expression, and hence the transition to the far-field pattern is difficult [64].

It is possible to represent the near field as a collection of plane waves traveling in all possible directions. The amplitude of these waves can be calculated if either the electric or the magnetic field is known over a suitably chosen aperture plane. The aperture distribution is then expressed as a Fourier integral and the radiated field calculated by relating the component exponential terms in the integral to plane waves [65], [66]. The approximations involved restrict the use of measurements to those made at appreciable distances from the antenna, although still in the near field. This complicates the picture somewhat because it is seldom possible to make the required measurements with a small near-field probe. A larger receiving or probing antenna is used, and the results so obtained are dependent upon the size of this receiving antenna and its directional pattern. For example, although diffraction theory suggests that sidelobe levels measured in the near-field are always higher than those in the far-field pattern, experimental measurements have shown that in the near field the measured sidelobe level is a function of the directivity of the receiving antenna as well as the separation between the antennas, and the measured sidelobe levels can be either larger or smaller than those in the far field [66], [67].

Kerns [68], [69] has developed a method for correcting for the presence of an arbitrary size probe antenna by characterizing an arbitrary two-antenna system by a plane-wave scattering matrix. For the case where multiple reflections between these two antennas can be neglected, and the scattering parameters of the probe are known, the measured near field data can be corrected for the perturbing effects of the probe.

This work is of major significance and has led to results of high accuracy [70]–[72]. The method is applicable in principle to any type antenna, however the amount of input data required may be large, and the suitability of its application will depend upon the precision of the far-field information that is desired.

Joy and Paris [73], [74] have established a sample spacing criterion for these near-field measurements made on a plane surface near an arbitrary antenna, and developed a near-field data minimization technique for reducing the required computational effort.

Another approach to the correction of the near field patterns is to express the radiated field as a Fourier series or a sum of cylindrical or spherical modes [75], [76]. The modes are the radially expanding fields which would arise in a radial transmission line, and the essential feature of the method is that the amplitudes and phase of these modes are deduced from measurements at any convenient distance from the antenna. In principle this procedure places no restriction on the distance at which measurements can be made. Thus there is no distinction between what can be regarded as near-field measurements and the measurement of the aperture distribution.

In very recent work, Leach and Paris [115] describe a method for determining the far-field pattern of an antenna from probe compensated near-field measurements over the surface of a right circular cylinder enclosing the antenna. The method is derived by first expanding both the field radiated by the antenna and the field radiated by the measurement probe when it is used as a transmitter, into cylindrical wave expansions. The Lorentz reciprocity theorem is then used to solve for the field radiated by the antenna from the probe output voltage. The antenna pattern can be determined independently of the characteristics of the measurement probe provided that certain calibration data for the probe are known, and a method for determining these data from the measured far field of the probe is described.

This method appears to be very interesting because of the ease with which the sampling can be accomplished.

C. Radio-Frequency Holography

Within the last few years there has been great interest in the possibility of reconstructing visible images from radio-frequency holograms [77], [78]. The literature is extensive and no attempt has been made to reference it in depth. Although not yet perfected, this work raises the possibility of obtaining visible images corresponding to the radiating near fields of radiating structures.

In order to solve the problem of reconstruction it is necessary to reproduce in a coherent optical system, a distribution of field which corresponds to the measured field of the antenna. Since a reproduction of a phase distribution is difficult at optical frequencies, a hologram of the antenna field may be taken as the initial distribution. This depends upon measurements of the interference pattern of the antenna field and a reference field. This reference field however, need not be introduced through space. It may be introduced in the transmission line system [79] and the measurements then reduce to simple probing over a surface in the vicinity of the antenna. In optical processing of a hologram of an antenna field, Kurochkin [80] has pointed out that the effect of the probe can be compensated by introducing into the proper plane of the optical system, a mask whose amplitude transmission coefficient varies in inverse proportion to the radiation pattern of the probe. However, the effect of the probe can be neglected if its angular dimension, as seen from the center of the antenna being tested, is an order of magnitude smaller than the width of the lobes of this antenna.

V. Near-Field Measurements in General

We have been concerned with the fundamentals involved in sampling the surface current and charge densities, and/or the field distribution, near these surfaces or over a given aperture. Much of the available literature on near-field measurements has been referenced. The serious experimentor will also find much valuable information in [39, ch. 14], which is concerned with the probing of fields to evaluate antenna ranges. In addition to those sources already referenced, some of the additional literature has been cataloged for convenience of the investigator who must apply these measurement techniques to a particular problem [81]–[114].

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