An Optically Linked Electric and Magnetic Field Sensor for Poynting Vector Measurements in the Near Fields of Radiating Sources

LANNY D. DRIVER AND MOTOHISA KANDA, FELLOW, IEEE

Abstract—A unique, single-element antenna measurement scheme that can simultaneously measure the electric, magnetic, and time-dependent Poynting vectors of electromagnetic (EM) fields is described. The electric and magnetic responses of the antenna sensor are separated by a 0°/180° hybrid junction. The resulting two RF voltages, along with relative phase and frequency information, are transmitted to a remotely located vector analyzer by a pair of well-matched fiber optic downlinks. The remote receiver measures and displays 1) the electric dipole response, 2) the magnetic loop response, and 3) the time phase difference between the two. This information is sufficient to determine the time-dependent Poynting vector. Both a theoretical analysis and a discussion of experimental measurements performed, which describe the capabilities and performance of a working prototype of the antenna measurement scheme, are presented. The results demonstrate that a three-axis (isotropic) version of this system could be used to measure the near fields of EM sources, as well as to completely describe the resultant flow of energy.

Key Words—Electric dipoles, electromagnetic compatibility, electromagnetic fields, magnetic loops, near-zone fields, Poynting vector.

I. INTRODUCTION

The rate at which energy traverses a surface in an electromagnetic (EM) field is indicated by the Poynting vector \( \mathbf{S} = \mathbf{E} \times \mathbf{H} \). It is apparent that the energy flow is normal to the direction of constant phase and in the direction of propagation. The mean value of the flow can be readily determined by

\[
\mathbf{S}_m = \frac{1}{2} \mathbf{E} \times \mathbf{H}^*.
\]  

In the far field region of an antenna radiating into an unbounded isotropic medium, the electromagnetic field is plane in nature, and the electric \( (E) \) and magnetic \( (H) \) field vectors lie in planes normal to the propagation vector. In this case, the magnetic field is propagated in the same direction, with the same velocity, and in exact time phase with the electric field. Their amplitudes differ by the intrinsic impedance factor, \( \zeta_0 \), of the medium. For free space

\[
\zeta_0 = \frac{\mu_0}{\sqrt{\varepsilon_0}} \equiv 120\pi \Omega
\]

where \( \mu_0 \) and \( \varepsilon_0 \) are, respectively, the permeability and the permittivity of free space. Therefore, the time averaged energy flux density \( (S_m) \) can be easily determined from the amplitude of either the electric field or the magnetic field as given by

\[
S_m = \frac{1}{2} E\mathbf{H} = \frac{1}{2} \zeta_0 \mathbf{E}^2 = \frac{\zeta_0}{2} H^2.
\]  

Many types of instruments have been developed for measuring either electric field or magnetic field strengths. These devices offer such special features as isotropic response, small size, high sensitivity, broad bandwidth, and minimal field perturbation. Typical electric field sensors (or probes) developed by the National Bureau of Standards (NBS) consist of three dipoles mounted orthogonally in three notches cut near the end of a plastic tube [1]-[3]. A high-frequency, low-barrier Schottky beam-lead diode is mounted at each dipole gap and serves as a detector. The probe unit employs high resistance lines to bring the detected dc voltages to an external metering unit. These lines also act as low-pass \( RC \) filters.

Isotropic probes designed primarily for measuring magnetic field intensities have also been developed at NBS which are based on the same principles and incorporate much of the same instrumentation used for the electric field probes [4],[5]. Here, three small orthogonal loops are used instead of short dipoles. Also, to provide a response that is flat over a wide frequency range, the \( Q \) of each loop antenna is reduced by means of resistive loading [4],[5].

As one moves from the far field region into the near-zone region of a transmitting antenna or a scatterer, the situation becomes considerably more complicated. For time-harmonic fields, the end points of the field vectors trace out polarization ellipses. Also, the electric and magnetic field vectors are not necessarily spatially orthogonal to each other or in time phase. Since the magnitude and phase of the more complex impedance is unknown, the true energy flux density cannot be determined simply from (3). Because the electric and magnetic field vectors are not necessarily perpendicular or in time phase, the Poynting vector (in the general case) lies on the surface of a cone with an end point on an ellipse. Thus, the time-dependent Poynting vector

\[
\mathbf{S} = \mathbf{E} \times \mathbf{H}
\]  

has to be evaluated from independent information of the electric and magnetic field components in order to describe the time energy flow. To date, all practical field strength probes measure either the electric field or the magnetic field separately.

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Manuscript received January 20, 1988; revised June 24, 1988.

The authors are with the Electromagnetic Fields Division, National Bureau of Standards, Boulder, CO 80303.

IEEE Log Number 8823345.
A new probe has been conceived and evaluated at NBS that measures the electric and magnetic fields simultaneously [6]. The probe consists of an electrically small, doubly-loaded loop that is capable of measuring the sum and difference of the induced electrical voltages that occur at opposite sides of the loop. Currents developed in the loads correspond to the electric dipole and magnetic loop responses. Across one load, the magnetic field response adds to the electric field response, while across the other load, the magnetic field response subtracts from the electric field response. A $0^\circ$/$180^\circ$ hybrid junction is used to separate the electric and magnetic field responses by providing the sum and difference of the induced voltages, as shown in Fig. 1. In an earlier publication [6], Kanda has provided a detailed theoretical analysis of this scheme. This paper presents detailed theoretical and experimental discussions of the character and measurement of the phase relationships which exist between the electric and magnetic components of EM fields, as well as the absolute magnitudes of these components. This phase information is crucial to the determination of not only the polarization ellipse of the field vectors (near-field environment) but also the time-dependent Poynting vector, which indicates the direction of energy flow.

II. THEORETICAL CONSIDERATION

Fig. 2 shows the geometry and coordinate system for a loop antenna with two gaps located at diametrically opposite points, $(\phi = 0, \pi)$, and loaded with equal impedances $(Z_L)$. It has been shown [6] that the currents $I(0)$ and $I(\pi)$ are given by

$$I(0) = 2\pi b E_0' \left( \frac{f_0 Y_0}{1 + 2 Y_0 Z_L} + \frac{f_1 Y_1}{1 + 2 Y_1 Z_L} \right)$$

and

$$I(\pi) = 2\pi b E_0' \left( \frac{f_0 Y_0}{1 + 2 Y_0 Z_L} - \frac{f_1 Y_1}{1 + 2 Y_1 Z_L} \right)$$

where $E_0'$ is the magnitude of an incident electric field, $f_0$ and $f_1$ are coefficients $(n = 0, 1)$ of a Fourier series expansion of the incident plane wave, and $Y_0$ and $Y_1$ are admittances for the magnetic loop current $(n = 0)$ and the electric dipole current $(n = 1)$.

For the quasi-static admittances, we get

$$Y_0 = \frac{1}{j\delta_0 k b \left( \frac{8 b}{a} - 2 \right)}$$

and

$$Y_1 = \frac{j2}{\delta_0} \frac{kb}{\left( \frac{8 b}{a} - 2 \right)}$$

$$f_0 = \frac{j k b}{2}$$

and

$$f_1 = \frac{1}{2}.$$
ships between the $\phi$ dependent gap currents and the corresponding output voltages (sum and difference) from the $0^\circ/180^\circ$ hybrid junction. The difference is simply the result of the sign and direction conventions being different for the two cases. As will be seen later in this paper, the practical implementation of the antenna yields a sum voltage that is proportional to the $E$-field and a difference voltage that is proportional to the $H$-field.

### III. Experimental Investigation

#### A. Background

Several prototype antennas had been fabricated and evaluated during a much earlier study of the feasibility of the basic EM field measurement scheme illustrated in Fig. 1 [6]. These devices were square in shape, and housed all of the signal processing circuitry required to produce two dc voltages, one being proportional to the $E$-field and the other proportional to the $H$-field. These dc voltages were easily transmitted to a remote metering unit by means of high resistance lines [7], [8], which neither perturbed nor responded to the EM field being measured.

The present application required that both frequency and phase information be maintained and conveyed to a remote RF vector analyzer. The high resistance lines, used previously, were unsuitable for this purpose because of their high RF loss and unpredictable phase distortion properties. Also, previous attempts to hardwire to such an antenna, e.g., using miniature doubly shielded $50\ \Omega$ coaxial cables, resulted in severe degradation of the $E$-field response. Consequently, the use of a pair of optical fiber downlinks seemed the most practical and effective means of relaying the two RF voltages, $V_E$ and $V_H$, or $V_E$ and $V_H$, to a remote receiver.

#### B. Fiber Optic Link Development

A pair of fiber optic links was designed for use with a 30 cm by 30 cm prototype $E$ and $H$ antenna. Each link consists of 1) an analog light emitting diode (LED) driver/modulator (optical transmitter), 2) a 10-m-long fiber optic cable, and 3) a p-i-n photodiode optical demodulator/receiver. The design made extensive use of modular components, e.g., a connectorized infrared LED module, a connectorized p-i-n photodiode module, and a transimpedance amplifier in an IC package. The photograph in Fig. 3 shows the retrofitted 30 cm by 30 cm antenna, the fiber optic cables, and the dual channel optical receiver. All of the circuitry required to produce the two RF voltages and to convert them to modulated optical carriers is contained within the antenna's lower half, which has a cross-sectional area of approximately 10 cm². Also, the rechargeable battery packs and regulators needed to supply the active devices are contained within the antenna interior. Fig. 4 shows a block diagram of the complete antenna system.

The circuitry comprising each analog LED optical transmitter, and its associated battery supply, is illustrated in Fig. 5 [9], [10]. Application of an RF voltage to the transmitter input causes a corresponding swing in the collector current of $Q_2$, thereby modulating the infrared optical power (at $\approx 850\ nm$) being emitted by the LED. Transistor $Q_2$ is quiescently biased.

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1 Figs. 4, 5, 7, and 8 contain various manufacturer part numbers. These numbers are included for completeness only, and in no way constitute any form of NBS endorsement.
such that it is never turned off or saturated. This pre-bias, along with the partially bypassed emitter resistance of Q2, compensate the LED, thereby extending the corner frequency \((f_c)\) well beyond its uncompensated value \((f_c \leq 50 \text{ MHz})\). This is referred to as "peaking" [11] and is illustrated in the spectrum analyzer traces of Fig. 6. Transistor Q1 provides temperature compensation and produces a "mirroring" effect on Q2. The values of \(R_1\), \(R_2\), and the dynamic resistance of the base-emitter junction of Q1 determine the input impedance of the transmitter. Resistors \(R_1\) and \(R_2\) were chosen to provide a nearly perfect 50 \(\Omega\) input impedance \((Z_{in} = 50.1 + j0.5)\) at 10 MHz with an RF input modulation signal of 0 dBm. This impedance remained nearly constant with RF drive level and frequency, maintaining an input VSWR \(\leq 1.10\) from 1 to 100 MHz at RF drive levels from -40 to 0 dBm. It is important that this impedance remain close to 50 \(\Omega\), as it constitutes the loading impedance for one output port of the 0°/180° hybrid junction. In addition to impedance matching, the values of \(R_1\) and \(R_2\) and those of \(R_3\) and \(R_4\) were selected for low quiescent (no signal) operating currents for Q1 and Q2. While this results in some reduction of dynamic range (= 6 dB) and gain (= 3 dB), it significantly reduces the current drain on the battery pack, thereby increasing operating time between recharges.

The optical demodulator/receiver is shown in the schematic diagram of Fig. 7. A p-i-n photodiode converts the modulated optical carrier to an RF current, which is fed into a transimpedance (TZ) amplifier. The amplified signal emerges as an RF voltage that is coherent with the field-induced voltage \((V_F\) or \(V_{in}\)) that produced it. Because the bandwidth of the p-i-n photodiode is strongly dependent on the reverse bias voltage applied to it, -18 Vdc was used to ensure a bandwidth of at least 160 MHz. This is in the same general range of that of the TZ amplifier used and was expected to be beyond the highest frequency of interest. The noise characteristics of this analog optical receiver are crucial to its performance, as they establish its sensitivity and that of the overall antenna measurement system. The receiver's major noise contributors are 1) the p-i-n photodiode, which generates shot noise that is proportional to the optical power level, and 2) the TZ amplifier, whose output noise is predominantly thermal in character, by virtue of the transresistance of the amplifier [12]. The optical carrier level used produces an RMS noise current spectral density, at the TZ amplifier input, which is essentially the same as that specified for the TZ amplifier \((\approx 2.5 \text{ pA/}\sqrt{\text{Hz}})\). By properly combining these two noise currents, multiplying the result by TZ, and converting the output voltage (across 50 \(\Omega\)) to power, a value of output noise power is obtained \((\approx 25 \times 10^{-18} \text{ W/Hz} \text{ or } -136 \text{ dBm/Hz})\). This noise power (normalized to 1 Hz) represents a threshold output level that establishes system sensitivity, i.e., only values of \(V_F\) and/or \(V_{in}\) that produce output levels above this threshold can be detected. Also, it should be noted that this threshold is directly proportional to system noise bandwidth. For example, with a noise bandwidth of 40 MHz, only EM fields that produce output levels greater than -60 dBm can be measured.

Prototypes of the optical transmitter and optical receiver are shown in the photograph of Fig. 8. The frequency response of the overall link is illustrated in the spectrum analyzer trace shown in Fig. 9. The -16 dB link gain is flat to \(\pm 0.5 \text{ dB}\) from 100 kHz to 120 MHz, and the 3 dB bandwidth is 30 kHz to 160 MHz. The -16 dB link gain corresponds to an output-to-input voltage ratio of approximately 160 \(\mu\text{V/mV}\). The
out of phase. The first two cases enabled selection of components, for the two channels, that yielded the closest gain and phase tracking. Case (3) simulated EM field induced signals in the antenna. In this manner the antenna and its associated optical links were thoroughly characterized over a frequency range of 1 to 40 MHz.

Having determined the amplitude and phase tracking correction factors, as functions of frequency and level, the antenna system was ready for evaluation in standard EM fields. A 3-m transverse electromagnetic (TEM) cell was used to provide known and controlled plane-wave EM fields [13], [14]. The retrofitted prototype antenna was placed in the center of the cell, oriented for maximum coupling with both the electric and magnetic components of the standard field. The two antenna outputs (from the optical transmitters) were connected to the dual optical receiver, located outside the cell, through the fiber optic cables. A highly sensitive vector analyzer was connected to the two outputs of the dual optical receiver, which measured and displayed the magnitudes of $E$, $H$, and the phase angle between them ($\theta_{E-H}$). This arrangement allowed measurements of frequency response, field strength response, and spatial-orientation response of the antenna system from 5 to 40 MHz in fields of 0.1-106 pW/cm$^2$. These ranges were selected on the basis of the operational performance of the TEM cell and on practical field density versus RF power considerations.

IV. RESULTS

The measured electric and magnetic responses of the antenna to EM fields between 5 and 40 MHz are shown in Fig. 10. Since these fields were established in the 3-m TEM cell and are therefore plane waves, it is both correct and convenient to express power density in equivalent plane-wave field strength (FS) units. Fig. 11 illustrates the field induced $E$ and $H$ antenna responses to field strength over a range of 1-106 pW/cm$^2$ (2-20 V/m). As expected, these responses increase linearly with FS, i.e., doubling the field strength increases the $E$ and $H$ voltage responses by 6 dB. The measured phase angle between the $E$ and $H$ voltages, $\theta_{E-H}$, was found to vary more with frequency than was theoretically predicted by (11) and (12), as shown in Fig. 12. This disagreement is suspected to be the result of using nonideal gap loading, i.e., loads that were not purely real (resistive). Unfortunately, time did not permit the resolution of this discrepancy.
The effects of gap loading on $V_E$, $V_H$, and $\theta_{E-H}$ were experimentally evaluated through the use of five pairs of miniature balun transformers. These baluns provided balance gap resistances of 50, 100, 200, 450, and 800 $\Omega$ when their unbalanced (single-ended) ports were terminated in 50 $\Omega$ via the port loading of the $0^\circ/180^\circ$ hybrid junction. Fig. 13 illustrates the dependence of the $E$ and $H$ responses on gap loading at 30 MHz. The effects of both gap loading and frequency on phase angle $\theta_{E-H}$ are demonstrated by the curves of Fig. 14. The theoretical curves of Figs. 15 and 16, produced from (11) and (12), indicate three cases of gap loading for which the phase angle $\theta_{E-H}$ is independent of frequency. Although the two extreme cases ($R_L \leq 0.1 \Omega$ and $R_L \geq 1 \text{ MQ}$) were not practically realizable, the moderate loading case ($R_L \approx 250 \Omega$) seemed easily obtainable. Unfortunately, the experimental curves of Fig. 14 indicate no such ideal value of gap loading. However, comparison with Fig. 15 reveals reasonably close agreement for gap resistances between 50 and 100 $\Omega$. More importantly, loading the two gaps with approximately 100 $\Omega$ produced 1) relatively constant
values of $\theta_{E-H}(15^\circ \pm 5^\circ)$ with frequency, and 2) nearly equal $E$ and $H$ responses—as shown in Fig. 13 and other curves (not shown) for frequencies between 1 and 40 MHz.

Other measurements made in the 3-m TEM cell revealed a system dynamic range that extended from 1 V/m to approximately 28 V/m for frequencies from 1 to 40 MHz. System sensitivity is established by the antenna response, the optical link gains, any optical fiber or connector losses, and the noise characteristics of the optical receiver. The upper FS limit is dictated by the linear operating range of the optical transmitter. This range could be increased by 6 dB to span 0.5–30 V/m by increasing the quiescent operating current of $Q_2$ from 18–40 mA. This doubles the optical transmitter gain at the expense of battery operating time (between charges) of the antenna battery packs. The typical system operating time (between battery recharges) is approximately 10 h.

The directivities of the $E$ and $H$ responses of the antenna were measured by rotating the antenna in the horizontal plane, and the results are shown in Fig. 17. The antenna was also positioned for 1) maximum $V_E$ and minimum $V_H$ and 2) maximum $V_H$ and minimum $V_E$. For these conditions the minimized voltage response was $\geq 35$ dB below that of the maximized voltage response. Earlier characterization measurements and simulated bench testing yield 40 dB of isolation between the $\Sigma(E)$ and $\Delta(H)$ channels of the antenna system. These cross-channel isolation characteristics are only slightly affected by frequency and field strength levels.

V. CONCLUSIONS

This paper has described a unique antenna configuration that has theoretically and experimentally demonstrated the capability of making simultaneous measurements of the electric, magnetic, and time-dependent Poynting vectors of EM fields. This dual-gap single-turn loop antenna system uses a $0^\circ/180^\circ$ hybrid junction to separate the electric and magnetic field responses, while maintaining both the fre-
Fig. 15. Theoretical phase angle as a function of frequency for several values of gap loading (resistance).

Fig. 16. Theoretical phase angle as a function of gap loading for 10, 30, and 40 MHz.

Fig. 17. Measured directivities of the $E$ and $H$ antenna responses at 30 MHz.

frequency and phase information needed to determine the Poynting vector. The probe design includes a pair of closely matched fiber optic downlinks that noninvasively transmit the sum and difference RF voltages from the hybrid junction to a remotely located vector analyzer. This receiver measures and displays 1) the electric dipole response, 2) the magnetic loop response, and 3) the time phase angle between the $E$ and $H$ responses. The phase angle can be used to provide the time-dependent Poynting vector, which, in turn, describes the time-dependent energy flow.

Plane-wave field measurements were made on a 30 cm by 30 cm prototype antenna over a dynamic range of 2–20 V/m over a frequency range of 1–40 MHz. The results were both repeatable and agreed favorably with theory. Considerable insight was gained relating gap loading to the $E$ response, $H$ response, the time phase angle between them, and frequency. Although no value of gap loading (using the balun transformers) was found that resulted in a frequency independent phase angle, $\theta_{E-H}$, adequate characterization and calibration can provide predictable measurement results. The fiber optic downlinks, designed for this antenna, can be used at frequencies approaching 200 MHz. Also, all components of the antenna are operable to at least 200 MHz. The physical size of the antenna would have to be considerably reduced, however, for proper operation at such increased frequencies.

The findings of this work indicate that the basic antenna design is capable of measuring the $E$ and $H$ field components and the Poynting vector of near-zone fields. An isotropic (three-element) version of the antenna would be required, as well as a metering unit capable of performing the necessary computations, for characterizing near-zone complex field distributions. Extending the frequency range would require a reduction in physical size of the antenna, which will be limited
by the size of the antenna battery packs required for the fiber optic transmitters.

There has been some speculation regarding other applications of the basic antenna scheme used, e.g., a totally passive antenna that would use fiber optic uplinking, voltage-sensitive birefringent crystals across the gaps, and fiber optic downlinking. It is quite possible that the vectorial addition and subtraction performed by the hybrid junction could be performed optically.

ACKNOWLEDGMENT

The authors express their thanks to S. D. Wytaske for his many contributions to the fabrication, characterization, and optimization of the fiber optic downlinks, and for his assistance in the evaluation and performance testing of the overall antenna system.

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