Adjustable compensation is achieved using the plunger arrangement of Fig. 5 in conjunction with the GP hole of Fig. 4 (normal GP thickness of 5 μm). With the plunger withdrawn approximately 100 μm away from the substrate, this hole overcompensates the transition. Adjustment of plunger-to-substrate distance X can produce a transition which is virtually transparent when examined by TDR. Measurements made on 25-mm-long 50-Ω microstrip lines on sapphire with properly compensated transitions at each end gave a total insertion loss rising from 0.1 dB at 1.0 GHz to 0.3 dB at 18 GHz. Fig. 6 shows a closeup view of the tab and plunger arrangement.

The launcher described is intended for precision laboratory measurement purposes. It is easily transferred between substrates and has proved very repeatable. Tab dimensions and the position of the GP hole are fairly critical along the line but symmetry about the line is not so important. There should be no air gap between the top of the substrate and the cable dielectric.

For the sapphire-substrate lines previously mentioned the optimum position of the plunger, dimension X in Fig. 5, is approximately 30 μm. Movement of the plunger 25 μm to either side of optimum produces reflections with $\Gamma \simeq 0.005$. The influence of the plunger ceases when X reaches 100 μm.

For use with alumina it has been found that for 0.514-mm-wide lines on 0.5-mm-thick substrates, a GP hole of 1.0-mm diameter is required, positioned over the plunger as in Fig. 5.

While the fully compensated transition is intended for a laboratory environment, it is suggested that the simpler version would be useful in production areas, particularly for automatic testing applications involving testing of individual circuit elements before integration into modules.

The fixed form of Fig. 3 is being used to examine circuits on substrates which have only the pattern on them. The GP used is a solid brass plate, suitably lapped, with 0.75-mm holes at various stations round the edge. Apparently the only penalty incurred in having a separate GP is a slight (10-percent) increase in losses above 10 GHz.

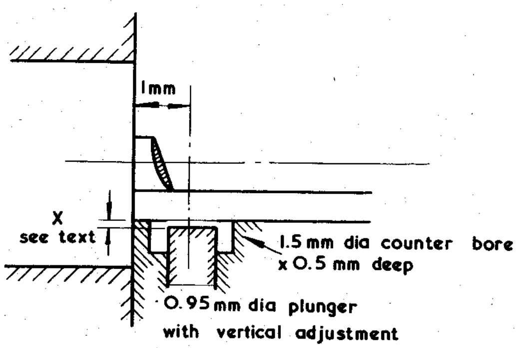

Fig. 5. Adjustable compensation plunger.

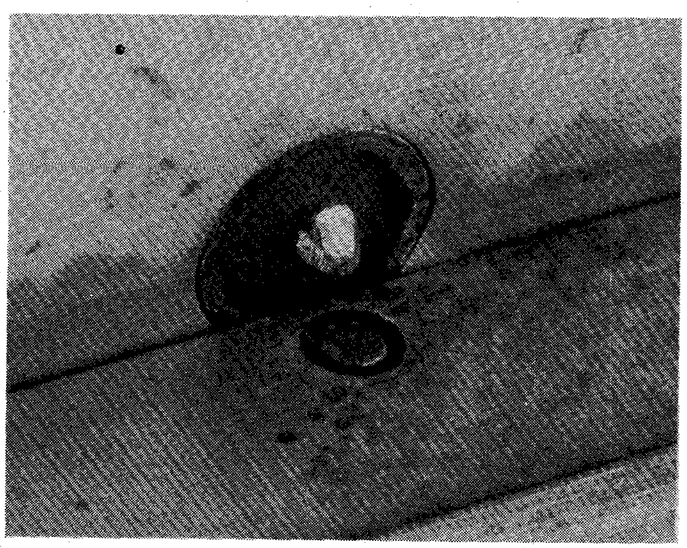
Fig. 6. Detailed view of tab and plunger.

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On an Automatic System for Simultaneous Measurement of Amplitude and Phase of Millimeter-Wave Fields

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***Abstract*—Suggestions are given for improving the vibrating dipole technique for measuring the phase and amplitude of millimeter-wave electric fields in free space. It is shown that the system can be simplified, at the same time reducing certain measurement errors and increasing the system's sensitivity and dynamic range. It is found that significant errors can result if the field being measured varies appreciably in amplitude, and/or if the phase does not vary linearly with position over the dipole's excursions.**

I. INTRODUCTION

Mathews and Stachera recently described a technique for measuring the amplitude and phase of electric field components in free space using a vibrating elemental electric dipole [1]. In another paper appearing in the same issue [2], they describe one method of automating this system for simultaneous amplitude and phase measurements. The chief merit of the vibrating dipole technique over other methods of modulating a scatterer is that the dipole can be made of the order of 1 mm long, permitting measurements of the electric field's fine structure at millimeter wavelengths.

The purpose of this short paper is to suggest three changes in this measurement system which will simplify it, eliminate certain measurement errors, and significantly increase the system's sensitivity and dynamic range.

II. PHASE SHIFTER LOCATION

Fig. 1 shows the modified automated system which operates on the same basic principles as those originally proposed in [2]. One important change is that the phase shifter has been transposed from point S in the antenna arm into the reference signal channel. This change has two main benefits. The first is that the procedure for tuning the antenna is considerably simpler. A mismatch in the antenna arm causes an unmodulated carrier signal to arrive at the detector via the information channel in addition to the unmodulated carrier arriving via the reference channel. Since the mismatch signal also coherently interacts with the modulated signal to produce an

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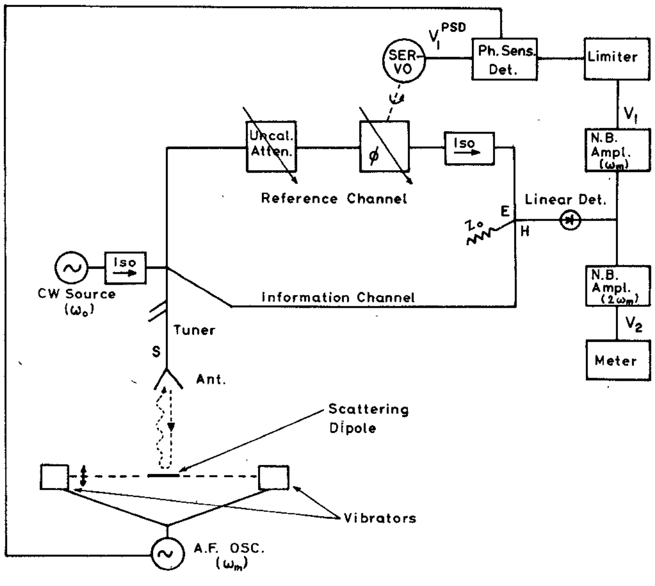

Fig. 1. Automated coherent detection system for measuring electric fields in free space using a vibrating dipole.

undesired output at the modulation frequency ω_m and its harmonics, the effect of the mismatch must be minimized. The most obvious way is to reduce the reflection from the antenna itself, but this is a very tedious process and virtually impossible to accomplish if the phase shifter is located at point S.

Another method uses balanced detectors. In [1, sec. V] it is demonstrated that a carrier signal entering the H-arm of the magic tee can be suppressed via the mixer's symmetry properties. This technique has been used on a number of occasions to reduce measurement errors and extend the dynamic range of coherent detection systems using amplitude modulation, e.g., see [3]–[6].

Tuning of the antenna arm is considerably simpler, and is constant if the variable phase shifter is not located at S, but rather, is located in the reference channel as shown in Fig. 1. This eliminates the need for the balanced detection system which has been replaced in Fig. 1 by a single mixer. However, there is the disadvantage that the range of variable phase shift is reduced by half.

III. SQUARE LAW VERSUS LINEAR DETECTION

The second benefit of placing the phase shifter in the reference channel is that the detected output can be made independent of the phase shifter's unavoidable and variable substitution loss. Mathews and Stachera [1], [2] assumed the detectors operate as square-law devices, but correctly noted that this is not required. Square-law operation gives an output amplitude which is in direct proportion to the amplitude of the reference channel signal as well as the information channel signal. But it has been shown (e.g., see [7]) that if the reference signal is sufficiently large, the detector characteristic becomes linear and simply responds to the envelope of the total input signal. The output then becomes independent of the amplitude of the reference channel signal. Thus the phase shifter's small but varying substitution loss has no influence on the output signal at ω_m or the harmonics of ω_m. In fact, the phase shifter can now be of an electrically controlled variety such as a Reggia–Spencer ferrite [8], [9], at least at microwave frequencies. The variation in the substitution loss of such devices is sometimes as large as 0.5–1.5 dB. This, in addition to their average substitution loss (≈ 1 dB), has precluded their use in precision measurement systems. Furthermore, phase changes as large as 10^3 electrical degrees are possible, greatly offsetting the reduced phase shift brought about by placing the shifter in the reference channel. Although such a device is reciprocal, a nonreciprocal shifter can also be used. Finally, the servo in Fig. 1 could be eliminated by connecting the output of the phase-sensitive detector (PSD) to an integrator followed by a dc amplifier which drives the electrically controlled phase shifter.

This independence of the output upon reference signal amplitude can extend over a considerable range. This is especially true for Schottky-diode detectors where ranges as large as 30 dB have been observed [7]. Furthermore, the system's greatest sensitivity generally lies within this range. Since the maximum information signal must be somewhat less than the reference signal, increasing the reference signal above the square-law region can increase both the upper and the lower ends of the dynamic range of the information signal.

As in all mechanical modulation methods, the modulation frequency is of the order of 100–200 Hz for the vibrating dipole scheme. This is unfortunate since the $1/f$ or "flicker" noise of the detector significantly reduces the system's sensitivity. Point contact devices characteristically have a large $1/f$ noise, making other lower noise devices such as the Schottky-barrier or backward diodes more attractive. Their use can increase the system sensitivity by 15–20 dB at low modulation frequencies [7]. The Schottky-barrier diode is particularly attractive since it has an IF impedance which is similar to that of a point contact. The small external forward bias usually required for Schottky-barrier diodes is not needed if the reference signal is large, as suggested previously. When rectified, this large reference signal is adequate to provide the necessary forward bias current.

However, backward diodes are not direct substitutes for point contacts. They have a very low IF impedance and must have no external bias or reverse self-bias. Consequently, a special IF amplifier which provides a proper IF impedance match and essentially zero dc input impedance is essential. The sensitivity near 100 Hz is very nearly the same as for a Schottky-barrier diode [7].

IV. SIMULTANEOUS PHASE AND AMPLITUDE MODULATION

Although not mentioned in the Mathews and Stachera papers [1], [2], it is also important to consider the presence of any amplitude modulation which occurs in addition to the phase modulation when using coherent detection systems. Furthermore, if the field's phase does not vary linearly with position (e.g., as $\exp(-jk_0z)$), then the spectrum of the phase modulated signal will be more complicated than that assumed. Thus the vibrating dipole technique must be used with caution when measuring fields which have a varying amplitude and nonuniform phase. Consider the present case where the phase modulation is accomplished by vibrating the scattering dipole at ω_m as shown in Fig. 1. The field intensity is generally not uniform over the dipole's excursions, and so the backscattered signal is also double sideband with carrier amplitude modulated to some degree. This would be particularly noticeable when measuring standing waves since the peak-to-peak excursion of the dipole is of the order of 120 electrical degrees to achieve a phase modulation index of 2 rad. The electric field in such a situation could vary several decibels, and the phase would not vary linearly with position.

In general, the backscattered modulated signal will have the complicated form

$$F_x^2[1 + f(t)] \exp \{ j[\omega_0 t + \phi(t)]\} \tag{1}$$

where $f(t)$ and $\phi(t)$ are the amplitude and phase modulation factors, respectively, and ω_0 is the angular radio frequency $F_x[= |F_x| \cdot \exp(j\phi)]$ is the component of the incident electric field which is parallel to the vibrating dipole at the midpoint of its displacement, and is the quantity to be measured in both amplitude and phase.

The envelope of (1) is then coherently detected by mixing with the unmodulated reference signal. Following an analysis similar to that used in [6], [7], the output of the linear detector is proportional to

$$|F_x|^2[1 + f(t)] \cos(2\phi + \phi(t)) \tag{2}$$

where we have assumed that the RF reference signal is much larger than the backscattered information channel signal [6], [7]. Here

2ϕ is the phase of the backscattered signal relative to the reference signal at the detector.

To gain some insight as to the measurement error involved when the amplitude of the field is not constant over the dipole's excursions, let

$$f(t) \cong m_1 \cos(\omega_m t + \psi_1) - m_2 \cos(2\omega_m t + \psi_2). \quad (3)$$

The amplitude modulation indices at the fundamental and second harmonics of the dipole's frequency of vibration ω_m are $m_{1,2}$, respectively. $\psi_{1,2}$ are phase angles relative to the corresponding phase modulated terms. For simplicity, we further assume that the field's phase changes linearly with dipole displacement, so that

$$\phi(t) = 2\phi_m \cos(\omega_m t) \quad (4)$$

where ϕ_m ($= 2\pi d/\lambda$) is the maximum phase change. Retaining only the first three terms

$$\cos(2\phi + \phi(t)) = J_0(2\phi_m)\cos(2\phi) - 2J_1(2\phi_m)\sin(2\phi)\cos(\omega_m t) - 2J_2(2\phi_m)\cos(2\phi)\cos(2\omega_m t) \quad (5)$$

where J_n are Bessel functions of the first kind and order n. Inserting (3) and (5) into (2), the resulting filtered outputs of the narrowband amplifiers centered at ω_m and $2\omega_m$ are

$$V_1 = A_1 | F_x |^2[(m_1/2)J_0(2\phi_m)\cos(2\phi)\cos(\omega_m t + \psi_1) - J_1(2\phi_m)\sin(2\phi)\cos(\omega_m t)] \quad (6)$$

and

$$V_2 = -A_2 | F_x |^2 \cos 2\phi[(m_2/2)J_0(2\phi_m)\cos(2\omega_m t + \psi_2) + J_2(2\phi_m)\cos(2\omega_m t)] \quad (7)$$

where the $A_{1,2}$ are constants which include the amplifier gains.

As shown in Fig. 1, V_1 is further processed by a limiter and a PSD. Ignoring the limiter for the moment, the dc output of the PSD is only that part of V_1 which is in phase with the audio frequency (AF) modulating signal $\cos \omega_m t$, or

$$\begin{aligned} V_1^{PSD} &= A_1' | F_x |^2[(m_1/2)J_0(2\phi_m)\cos(2\phi)\cos(\psi_1) - J_1(2\phi_m)\sin(2\phi)] \\ &= A_1' | F_x |^2\{[(m_1/2)J_0(2\phi_m)\cos\psi_1]^2 + J_1^2(2\phi_m)\} \cdot \sin\left[2\phi - \tan^{-1}\left(\frac{m_1 J_0(2\phi_m)\cos\psi_1}{2J_1(2\phi_m)}\right)\right] \end{aligned} \quad (8)$$

where A_1' is a constant which includes the gain of the PSD. This signal is used to drive the servo which seeks a zero crossing of (8) by changing the phase, ϕ. This occurs at

$$2\phi_0 = \tan^{-1}\left(\frac{m_1 J_0(2\phi_m)\cos\psi_1}{2J_1(2\phi_m)}\right). \quad (9)$$

The zero crossings and subsequent phase error are not affected by the limiter which simply prevents the PSD from being overdriven.

If the amplitude modulation is very small, or if $2\phi_m \simeq 2.4$ so that $J_0 \cong 0$, these zero crossings are shifted only slightly from $\pm n\pi$ where the crossings would have been if there were no amplitude modulation. Furthermore, if $2\phi_m \cong 2.4$ the amplitude error in V_2 given by (7) is also minimized, but this condition may be difficult to maintain if the phase variation is nonlinear.

If amplitude modulation is not considered, the dipole's excursion would normally be adjusted so that $J_2(2\phi_m)$ is maximum ($2\phi_m = 3.1$). Then $J_0(3.1) = -0.292$, $J_1(3.1) = 0.3$, and taking $\psi_1 = 0$ in (9), the phase error is $2\phi_0 = -8.3°$ if $m_1 = 0.3$. Also, note that the peak-to-peak phase change is almost π. Further, such a large dipole excursion can lead to significant second harmonic amplitude modulation, i.e., m_2. In turn, this leads to an amplitude measurement error using V_2.

These amplitude and phase errors can also be reduced by interchanging the roles of V_1 and V_2, at least in cases where the phase changes linearly. To show this, V_2 is coherently compared with $\cos(2\omega_m t)$ by a PSD detector. Some PSD's (e.g., a lock-in amplifier) are capable of doing this by using the fundamental $\cos \omega_m t$ reference signal. Then the dc output of this PSD is

$$V_2^{PSD} = A_2' | F_x |^2 \cos(2\phi)[(m_2/2)J_0(2\phi_m)\cos\psi_2 + J_2(2\phi_m)]. \quad (10)$$

When this signal is fed to the servo, the system will stabilize so that $\cos 2\phi_0 = 0$, or $2\phi_0 = \pm(2n+1)\pi/2$ regardless of the amplitude modulation. The servo position is therefore a direct indicator of the correct phase.

The relative field amplitude is indicated by V_1. For $\cos 2\phi_0 = 0$, (6) becomes

$$V_1 = -A_1 | F_x |^2 J_1(2\phi_m)\cos\omega_m t$$

which is also independent of the incidental amplitude modulation.

An additional benefit of using V_1 to measure amplitude and V_2 to control the phase via the servo is that the phase modulation index $2\phi_m$ can be somewhat less than when their roles are reversed. For example, $J_1(2\phi_m)$ is maximum for $2\phi_m = 1.84$ which is significantly less than the value of 3.1 which is required to make $J_2(2\phi_m)$ maximum.

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Time-Domain Measurement of Loss and Dispersion

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***Abstract*—Time-domain metrology (TDM) techniques are applied to determine loss and dispersion in microstrip and coaxial cable for possible future use in interconnections in the frequency range of 0.4–10 GHz. After a brief presentation of the method, results are given for microstrip, RG/U 58, and RG/U 188A coaxial cable. Good agreement is obtained between measured, computed, and published values. Major advantages of the technique are that unwanted multiple reflections can be excluded from the measurement time window, and errors from interface discontinuities can be evaluated and removed from the final result.**

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