1. Introduction

In this paper, we review the system design, antenna structures, and optoelectronic components that can be used to accurately measure electromagnetic (EM) fields at frequencies up to 100 GHz, in support of the technologies of telecommunications, materials processing, health services, and defense. In addition, electromagnetic compatibility and interference (EMC/EMI) in the electronics that support these and other technologies can result in reduced performance and productivity, and in hazardous conditions. Sensors with broad, flat frequency responses are important tools for accurately determining EM fields in EMC/EMI applications, especially when transient or pulsed fields are involved.

EM-field sensors with photonic links reduce or eliminate some of the inaccuracies and systematic errors that affect measurement techniques using conventional EM-field sensors. They provide electrical isolation, which eliminates ground loops and common-mode electrical pickup between the sensor head and electronics module. The optical fibers and dielectric components produce minimal field distortion. In addition, they can preserve both the phase and amplitude of high-frequency fields, with good fidelity and low losses. These are characteristics that also make it possible for us to reduce the uncertainties in antenna calibrations performed at our open-area test site, by filtering out signals from the ambient. The benefits of an optical link were recognized by Hertz [1893] more than a hundred years ago, when he used a telescope to observe sparks in the gap of a loop antenna, during the landmark experiments that verified the wave nature of electromagnetic radiation. The development of optical-fiber and optoelectronic components for the telecommunications industry has made it possible to implement photonic sensors that are accurate and convenient to use.

We focus on passive systems, as shown in Figure 1, that require little or no power at the sensor head, and produce minimal perturbation of the measured field. Light from a laser is transmitted over an uplink optical fiber to an electro-optic (EO) modulator in the sensor head. The light intensity is modulated by the voltage from a
Figure 1. A block diagram of a passive, electro-optic electromagnetic-field sensor with fiber-optic links.

small antenna. The modulated light is transmitted back, through a downlink fiber, to the receiving-and-control electronics. There, the optical signal is detected, amplified, and processed to give an output that is compatible with the inputs for spectrum and network analyzers or oscilloscopes. A thermo-optic technique, which does not preserve the temporal phase of the field, is discussed by Randa et al. [1992].

2. System concept and analysis
2.1 Signal

An analytical model allows us to predict system performance, and the sensitivity to various design parameters. For this, we use the signal currents, $i_s$, out of the receiver photodiodes. For channel A, the signal current can be written in terms of a transfer function,

$$i_s^A = P_o S_\phi^A S_r^A S_m^A,$$

(1)

where $P_o$ is the optical power from the laser; $S_\phi^A$ is the attenuation through the channel due to losses in fibers, couplers, connectors; the excess or insertion loss of the modulator, $S_r^A$, is the conversion efficiency of the photodiode, including avalanche gain, if present; and $S_m^A$ is the modulator transfer function for the
channel. The modulators' transfer functions are generally nonlinear. However, a quiescent or bias point can be chosen, so that operation over an adequate linear and dynamic range is achieved for small input signals. When this is true, the light intensity out of the modulator can be expressed in a linear form by

\[ S_m = S_{bias} + \frac{\partial S_m}{\partial V} V(t) = S_{bias} + \frac{\partial S_m}{\partial V} S_a E_{rf}(t), \]  

which consists of a dc component, \( S_{bias} \), that is modulated by the time-varying voltage on the modulator electrodes, \( V(t) \). The voltage on the modulator electrodes is given by the electric field, \( E_{rf}(t) \), and the antenna factor, \( S_a \). We consider here only modulators that have two output channels, with the ratio of the optical power in the channels being a function of the applied voltage. This excludes the use of electro-absorption modulators, such as those described by Ido et al. [1997] and Liao et al. [1997], which are of considerable interest for telecommunications. The complementary outputs of channels A and B are characterized by

\[ S_m^A + S_m^B = 1 \quad \text{and} \quad \frac{\partial S_m^B}{\partial V} = -\frac{\partial S_m^A}{\partial V}. \]  

As is conventionally done, we use the difference in the signals from the two channels (balanced detection), and then normalize it by the sum of the two signals, to obtain

\[ \frac{i_s^A - i_s^B}{i_s^A + i_s^B} = P_o \left( S_o^A S_r^A S_{bias} - S_o^B S_r^B S_{bias} \right) + P_o \left( S_o^A S_r^A + S_o^B S_r^B \right) \frac{\partial S_m^A}{\partial V} S_a E_{rf}(t) \]  

By either adjusting the gains in the photoreceivers or attenuating the light in one of the fibers, we can set \( S_o^A S_r^A = S_o^B S_r^B = S_o S_r \). Equation (4) then reduces to

\[ \frac{i_s^A - i_s^B}{i_s^A + i_s^B} = \left( S_{bias}^A - S_{bias}^B \right) + 2 \frac{\partial S_m^A}{\partial V} S_a E_{rf}(t), \]  

where we have used the results of Equations (2) and (3) to set the denominator equal to 1. The received signal strength is no longer dependent on the received optical power, nor subject to fluctuations in it due to drift or relative intensity noise (RIN) in the laser. The \( S_{bias}^A - S_{bias}^B \) term is ideally a dc, or slowly varying offset, which can be capacitively blocked from the RF circuitry. However, if the modulator bias point is fluctuating rapidly, it will result in a time-dependent signal that may have frequency components within the detection bandwidth. Although modulator designs
having a stable, optimum bias point at $S^A_{bias} = S^B_{bias}$ and zero drift would eliminate this term, most modulators drift over time (primarily due to temperature changes), and some sort of active bias control is required for practical applications outside a laboratory. We have assumed there is no difference in arrival times at the inputs to the sum and difference circuits for signals from channels A and B, or between the two signals going into the divide circuit. If this is not the case, there will be incomplete cancellation of the fluctuations in the laser power. Therefore, it is important to keep the optical cables close to the same lengths, and to provide tunable delay lines, in order to make fine adjustments to the timing.

2.2 Noise

Noise currents accompany and degrade the signals. They set a lower limit on the detectable fields, and to a large part, determine the useable dynamic range for the system. There are several noise-generating mechanisms, which produce uncorrelated noise currents that are normally characterized by Gaussian distributions [Hentschel, 1988]. The quadratic addition of these currents results in a mean-square-noise current given by

$$i_n^2 = B_n \sum i_j^2 = B_n \left[ \frac{4kT}{R_t} + 2e (rP_d + I_d) M^2 F(M) \right],$$

(6)

where $B_n$ is the detection bandwidth, $k$ is the Boltzmann constant, $R_t$ is the resistance that terminates the photodiode, $T$ is the absolute temperature of $R_t$, $e$ is the charge of an electron, $r$ is the responsivity of the photodetector, $P_d$ is the optical power onto the photodetector, $I_d$ is the photodiode dark current, $M$ is the avalanche gain, and $F(M)$ is the excess noise factor for an avalanche photodiode (for a PIN diode, $F(M) = M = 1$ and $S_r = r$). Although relative intensity noise is often a major contributor to the noise in optical-fiber systems with fairly high optical power, we have not included it here as a noise term, because it cancels out in the signal detection and normalization we have chosen. We have also left out contributions due to modal noise, because it is minimal when using single-mode fiber. Noise due to mixing of polarization states in the uplink fiber can be converted into common-mode relative intensity noise by inserting a polarizer directly in front of the modulator. Polarization mixing in the downlink fibers can be a problem if the photodetectors exhibit substantial polarization sensitivity.

When the noise currents are added to the signal currents, Equation (4) can be written as
where we have assumed that the noise currents in each channel are equal, and have added them quadratically. The noise terms are normalized by the received optical power. The normalized noise term in the denominator is much less than 1, and contributes insignificantly to the uncertainty of the received signal. For a well-designed system, shot noise, which is given by $2erP_dB_n$, is the dominant contributor to the total noise. Thus, the noise contribution to the normalized signal amplitude is proportional to $1/\sqrt{P_o}$.

A figure-of-merit for a sensor system’s performance is the noise-equivalent field, $E_{neq}$. This is defined as the field level which would produce a signal just equal to the noise in a 1 Hz bandwidth. We calculate this from Equation (7) by assuming that the shot-noise term is dominant, that the denominator is nearly 1, and that the $S_{bias}$ terms are equal and changing slowly enough that they make no significant contribution to the noise. Then, using a 1 Hz bandwidth, we equate the remaining terms in the numerator, to obtain

$$ E_{neq} = \frac{e}{\sqrt{2PoSoSr}} \left( S_a \frac{\partial S_m}{\partial V} \right)^{-1}. \quad (8) $$

3. Antennas

If the EM field is sufficiently large, it can produce adequate optical phase shifts directly in an EO crystal, without the gain from an antenna. Thus, all metallic or conducting components can be eliminated from the sensor head, and the modulator design is simplified. In addition, the frequency response is broadened, because it is no longer limited by the electrical characteristics of the antenna and by the modulator’s electrode capacitance.

For most applications the gain of an antenna is required for adequate system sensitivity. For many measurements, and especially for pulses, an antenna with a flat, broad-band frequency response is desirable. The transfer function, $S_a(f)$, of a linear antenna is defined as the ratio of the received voltage, $V_r(f)$, to the incident electric-field component, $E_i(f)$, parallel to the antenna, and is given by
\[ S_a(f) = \frac{V_i(f)}{E_i(f)} = \frac{h_e(f)Z_L(f)}{Z_L(f) + Z_0(f)}, \]  

where \( h_e(f) \) is the antenna’s effective length, \( Z_0(f) \) is the input impedance of the antenna, \( Z_L(f) \) is the load impedance, and \( f \) is the frequency. We have suppressed the \( \exp(j2\pi ft) \) time dependence.

### 3.1 Electrically short dipole

For an electrically short dipole antenna \((\beta h \leq 0.5)\), the effective length and input impedance are given by King [1956] to first order as

\[ h_e = \frac{h(\Omega - 1)}{2(\Omega - 2 + \ln 4)} \]  

and

\[ Z_0(f) = \frac{\xi_0}{2\pi \beta h} \left[ \frac{(\beta h)^3}{3} - j(\Omega - 2 - \ln 4) \right], \]  

where \( h \) is the physical half-length of the dipole antenna, \( \beta = 2\pi/\lambda_0 \) is the wave number for the free-space wavelength \( \lambda_0 \), \( \xi_0 \) is the free-space wave impedance \((120\pi \text{ ohms})\), \( \Omega = 2\ln(2h/a) \) is the antenna thickness factor, and \( a \) is the antenna radius. The EO modulators that we use provide an almost purely capacitive load across the antenna terminals, which is given by

\[ Z_L(f) = -j \frac{1}{2\pi fC_L}, \]  

where the capacitance, \( C_L \), is a few \((1 \text{ to } 10) \) picofarads. We neglect the very small real part of Equation (11), and combine Equations (9) through (12) to obtain a frequency-independent transfer function of

\[ S_a(f) = \frac{h\alpha}{1 + C_L/C_a}, \]  

where the antenna capacitance, \( C_a \), is given by
Electrically short, thin, metal dipoles work well, as long as the power spectrum of the incident radiation contains little energy at wavelengths $\lambda$ greater than $h/3$, and the field is strong enough to produce adequate drive voltage across the EO modulator. Longer dipole lengths provide greater signal strength, but become resonant if the field contains spectral energy at shorter wavelengths.

### 3.2 Resistively tapered dipole

As a means of overcoming the natural dipole resonance, a traveling-wave dipole antenna, realized by the continuous resistive tapering of each element [Wu and King, 1965; Kanda, 1978; Kanda, 1986; Kanda and Driver, 1987], can be used. If the internal impedance per unit length, $Z(z)$, is given as a function of the axial coordinate, $z$, by

$$Z(z) = \frac{60\psi}{h - |z|},$$

then the current distribution, $I_x(z)$, along the linear antenna is that of a traveling wave, which does not exhibit resonant behavior. That is,

$$I_x(z) = \frac{V_0}{60\psi (1 - j/\beta h)} \left[ 1 - \frac{|z|}{h} \right] e^{-j\beta |z|},$$

where

$$= 2 \left[ \sinh^{-1} \frac{h}{a} - C(2\beta a, 2\beta h) - jS(2\beta a, 2\beta h) \right] + \frac{j}{\beta h} \left( 1 - e^{j\beta h} \right),$$

and $C(x, y)$ and $S(x, y)$ are the generalized cosine and sine integrals. The remaining symbols were defined previously. The effective length of a resistively tapered antenna exhibits no resonant characteristics, as shown in Figure 2. The frequency response of one of our sensors, consisting of a 0.15 m resistively tapered dipole with an EO modulator having about 3 pF capacitance, is shown in Figure 3.
Figure 2. The frequency response of dipole antennas, with and without resistively tapered elements.

Except for crystal resonances between 1 MHz and 3 MHz, the response is independent of frequency below 100 MHz. The resonances are acoustic modes driven by the piezoelectric coefficient of the Bi$_4$Ge$_3$O$_{12}$ crystal. These can be almost eliminated by a careful choice of mounting materials and physical geometry of the modulator chip. Between 100 MHz and 2 GHz, the response shows a slight dip and then rise, which is attributed to a small inductance in the leads between the dipole elements and the modulator electrodes. The ragged signal above 100 MHz is due to mode mixing in the reverberation chamber used for the measurements. At the high-frequency end, the response falls off due to the reduced effective length of the dipole.

4. Modulators

The measurement accuracy of a sensor system depends partially on the properties of the EO modulators. The basic principles and designs for EO modulators are described by Yariv and Yeh [1984]. The modulators of interest for EM-field measurements are based on the EO effect exhibited by some crystals, in which the optical index of refraction, and, hence, the propagation constant for light, are a linear function of an applied electric field. This effect is used to make a variety
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\frac{C}{C_0}
\]

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Figure 3. The frequency response of an electromagnetic-field sensor with a \( \text{Bi}_4\text{Ge}_3\text{O}_{12} \) Pockels-cell modulator, and a resistively tapered dipole (each element being 75 mm long).

of interferometers, directional couplers, and Pockels-cell polarimetric modulators. All of these designs can be implemented using optical-fiber components, together with waveguide devices fabricated in \( \text{LiNbO}_3 \) [Alferness, 1990; Wong, 1991], by epitaxial growth of GaAs [Hunsperger, 1991], or in polymer films [Park, 1997]. Our discussion focuses on those modulator characteristics—such as sensitivity, dynamic range, and stability—that are germane to their use in EM-field sensors. Modulators fabricated as waveguide devices are considerably smaller than their counterparts fabricated from bulk crystals and discrete coupling lenses. Also, their sensitivities are approximately 100 times greater than for corresponding bulk devices, because the closely spaced electrodes that are achievable with photo-lithographic techniques produce much stronger electric fields in the waveguide regions of the crystals. For these reasons, and because of their greater mechanical stability, we prefer waveguide modulators for most applications.

4.1 Sensitivity and frequency response

As with many devices, EO modulators exhibit a tradeoff between sensitivity and bandwidth. The sensitivity is proportional to the length of the region in the crystal for interaction between the light and the electric field, but the frequency
response falls off with increased interaction and electrode length. For lumped-element electrodes, the frequency response is primarily limited by the $RC$ time constant associated with the capacitance of electrodes [Becker, 1985]. When fabricated in LiNbO$_3$, modulators with lumped-element electrodes typically have maximum frequency responses in the 2 to 5 GHz range. The frequency response of a modulator is also limited by the overlap in time between the passage of light through the modulator and the quasi-stationary electric-field condition [Yariv and Yeh, 1984]. Traveling-wave electrodes extend the frequency range, by having transmission-line characteristics that more closely match the phase velocity of the light and the electric signal [Becker, 1984]. Modulators with traveling-wave electrodes have been reported with frequency responses up to 40 GHz in LiNbO$_3$ [Gopalakrishnan et al., 1992], up to 50 GHz in GaAs [Walker, 1993], and to above 100 GHz for polymers [Chen et al., 1997]. A modulator with small antennas as electrodes that were phase-matched to an incident millimeter-wave field was reported by [Bridges et al., 1991] to have a frequency response above 90 GHz.

4.2 Bias-point control

In order to obtain reproducible measurements, it is important for the modulator’s transfer characteristics to remain stable, and not to drift with changes in the environment. Although some modulators can be fabricated with a correct bias built in, variations in crystal uniformity and processing parameters make it difficult to reproducibly achieve the optimum point. Even with post-processing steps, such as laser trimming [Bulmer et al., 1991] to optimize the passive bias, it usually will still drift with changes in temperature. In LiNbO$_3$, drift arises from voltages induced across the crystal by temperature change and the pyroelectric effect. Although use of $x$-cut crystals and an electrical connection between its $z$ faces is very effective in minimizing the drift [Skeath et al., 1986; Greenblatt et al., 1995], the stability achieved still does not appear to meet the requirements for a precision measurement system. Also, the additional processing and the piecewise trimming of parameters add substantially to the cost of modulators for sensor applications.

The addition of a feedback loop, to actively control the bias point, may be a better solution. The error signals needed for the feedback control can be obtained from an imbalance in the dc outputs from the modulator, or by minimizing the dominant, second harmonic from a low-frequency dither signal applied to the control system. The latter approach has the benefit of not being affected by changes in the optical intensities resulting from disconnecting and reconnecting the fiber couplers. The bias point of the modulators can be adjusted by changing the polarization state or wavelength of the laser light at the input, or by applying a bias voltage to electrodes on the waveguide. The bias voltage can be supplied by power from a separate uplink laser to a small photocell array at the sensor head. However, drift of the bias under applied dc voltage is a problem in LiNbO$_3$, especially when SiO$_2$ buffer layers are used [Nagata et al., 1997], and a method to correct for this
may be necessary. The best technique depends on the specific modulator design, and whether miniature photovoltaic arrays are compatible with the intended use. With the necessary components for the photocell approach now commercially available, this approach may prove to be the least expensive and easiest to implement for many applications, even with the added complexity caused by the dc drift.

4.3 Dynamic range

The useful dynamic range of a sensor depends on the system’s noise floor, the linearity of the modulator transfer function, and the intended application. The basic nonlinear response of modulators leads to the generation of harmonics, which become more pronounced as the signal levels increase. We distinguish between two cases. If the input is a single-frequency, narrow-band signal, a spectrum analyzer can be used to filter out the harmonics and give a calibrated response at the fundamental input frequency. In this case, we use the 1 dB-compression point for the maximum allowable signal. We choose the minimum signal as the point where the response is 3 dB above the shot noise, in a 1 kHz bandwidth, for 1 mW of optical power onto a typical detector.

If the source is broadband, with many frequencies present, intermodulation products generated by the harmonics will limit the useful upper signal. Pulse shapes can become distorted. In this case, we use a standard analysis [Gaskell, 1958] to calculate the harmonic coefficients, and choose the minimum and maximum signals as the points where the power in the fundamental and in the dominant harmonic rise 3 dB above the noise floor, for a 1 GHz detection bandwidth. The choice of detection bandwidth is arbitrary, and would most likely be different in each application. However, the bandwidths we chose are reasonable, and allow us to compare the dynamic ranges for the different modulators under somewhat realistic conditions. The bias-point position also has considerable effect on the harmonic content. For a Mach-Zehnder interferometric modulator, a drift in the bias point of 5° away from the ideal results in a substantial increase in the second harmonic, and reduces the dynamic range by more than 10 dB [Howerton et al., 1990].

4.4 Modulators with complementary outputs

Modulators with complementary outputs are the best choice for accurate and repeatable EM-field measurements. As shown in Section 2, the normalized, balanced detection technique can give a 6 dB increase in the signal power, while reducing the common-mode noise. The complementary outputs can also provide signals for the bias-control feedback circuits. Figure 4 shows the principal guided-wave modulators and the additional components necessary for obtaining the complementary outputs. We present the analysis of a Mach-Zehnder modulator to obtain its sensitivity, dynamic range, and bias-control characteristics. The results are summarized,
together with the results of similar analyses for the other modulators, in Table 1. In Figure 5, we plot all the transfer functions over a range of normalized input voltages. The choice of modulator really depends on the ease of implementing the feedback control for the bias point, the sensitivity needed for the application, and costs.

4.4.1 Mach-Zehnder interferometer (MZI)

The transfer function for a two-beam interferometer, such as a Mach-Zehnder or Michelson interferometer, is relatively simple. It is discussed in standard texts [Hecht, 1990], and can be written as

\[ S_{MZ} = \frac{I}{I_0} = \cos^2 \left( \frac{\pi V_n}{2} + \frac{\Phi_{bias}}{2} \right), \]  

(19)

where we have used a normalized input voltage, defined as \( V_n = \frac{V_L}{V_\pi} \). \( V_L \) is the voltage on the electrodes, and \( V_\pi \) is the voltage required to produce a phase shift of...
Figure 5. Plots of transfer functions for several electro-optical modulators.

π between the two channels. Actual modulators do not exhibit full modulation depths, and it is convenient to characterize them by their extinction ratio $I_{\text{min}}/I_{\text{max}}$, for the light intensity in an output channel. Extinction ratios of -20 dB are common. We define the modulator's sensitivity as the derivative of the transfer function, evaluated at the bias point. A MZI modulator, biased at $S^{A}_{\text{bias}} = S^{B}_{\text{bias}}$ yields a maximum sensitivity given by

$$\frac{\partial S_{\text{MZI}}}{\partial V_L} = \frac{\pi}{2V_{\pi}}. \quad (20)$$

The asymmetry in the modulator shown in Figure 4 is calculated to produce a difference in the two optical paths of $\lambda/4$, as needed for an optimum passive bias point. Changing the bias point for this type of modulator, by changing the laser wavelength, would require a tuning range that is too large for practical application. Furthermore, the coupling ratio in the waveguide output coupler is also subject to temperature-related drift. In some commercial products, intended for the analog cable-television market, voltages from a miniature photovoltaic array are used to offset the drift in both the interferometer and the coupler sections. Such a package would not be desirable for most E-field sensor systems, because considerable electronic circuitry is involved.
Figure 6 shows the results of the harmonic analysis for an MZI modulator, biased at the optimum point, with input signals of frequency $\omega$ and various amplitudes. Because of the symmetry of the transfer curve, only the odd harmonics are present, and the power at $3\omega$ is dominant. The dynamic ranges are shown for signal minimums that are 3 dB above the shot noise in 1 kHz and 1 GHz detection bandwidths, with 1 mW of optical power at the detector. The dynamic range of 110 dB for the narrow-band signal is reduced to 46 dB when the bandwidth is increased to 1 GHz. These results are included in Table 1. The dynamic ranges for practical systems will be degraded by an increase in the noise due to other sources, and due to any drift in the bias points. When adjusted to a 1 kHz bandwidth, a measured dynamic range of 100 dB has been reported in the literature by Kuwabara et al. [1992], for an MZI modulator.
4.4.2. Michelson interferometer

The transfer function and sensitivity for the Michelson interferometric modulator are also given by Equations (19) and (20). Consequently, its linearity and dynamic range have the same characteristics as those of the MZI modulator. The input and output fibers both attach to the same end of the modulator chip, and the light is reflected back through a 3 dB coupler. Thus, the modulator can be conveniently located at the end of a handle that houses the ancillary optical components. A Michelson modulator, configured with a Faraday rotator and polarization splitter to remove the downlink signal from the uplink fiber has been incorporated into a standard reference antenna by Masterson et al. [1996a]. In this application, separate phase modulators, with one end mirrored, were attached to fibers from a 3 dB optical-fiber coupler. The benefits of this design are that an optical-fiber coupler is more stable than one fabricated as a waveguide on a chip; the modulator chips can be placed back to back with conducting adhesive, in order to further reduce their sensitivity to temperature changes; and a 0.5 mm difference in the fiber lengths for the two arms enables us to maintain the optimum bias point by controlling the laser wavelength. The feedback control has worked well, even though the decreased sensitivity to thermal drift in the modulator chip has been partially offset by fluctuations in the optical-path lengths in the fibers of the coupler. The reflection of the light back through the modulator doubles the sensitivity over that of an MZI modulator with the same electrode lengths. With the 4 mm-long electrodes, a value of 1.8 V was obtained for $V_n$.

4.4.3 Pockels cell

A Pockels-cell polarimetric modulator is basically a two-beam device that also has the same sinusoidal transfer function and dynamic range as the MZI modulator. Although traditionally fabricated from bulk crystals, it can also be implemented by titanium-indiffusion processing of LiNbO$_3$ [Alférness, 1990], which produces...
waveguide channels that support the propagation of both polarization states. A relatively simple modulator can be assembled by orienting a polarizing uplink fiber so that it launches light with equal amplitudes into both polarization eigenstates for the LiNbO$_3$ waveguide. An optical-fiber polarization splitter on the output gives the complementary signals. The bias point can be adjusted by voltages from a photovoltaic array. For a 1 cm-long waveguide, it could also be adjusted by tuning a 1.3 µm-wavelength laser over a range of approximately 3 nm. The Pockels-cell modulator can also be assembled as a reflection modulator, using Faraday rotators and polarization splitters. An attractive variation to this method, proposed by Enokiwhara et al. [1987], is to use polarization-rotated reflection (PRR). This allows the use of a single polarization-maintaining fiber to the modulator for both the uplink and downlink signals. It can simultaneously reduce noise at the receiver, due to phase fluctuations in the polarization state. However, a number of ancillary components are required at the optical-receiver module.

### 4.4.4 Fabry-Perot interferometer

Fabry-Perot (FP) etalons can also be used as modulators for EM-field measurements [Masterson et al., 1996b]. The transfer function is described by the Airy function [Hecht, 1990], and is shown in Figure 5. The Airy function and the modulator sensitivity, given in Table 1, depend strongly on the fringe visibility, $F$, given as

$$ F = \frac{4R}{(1 - R)^2}, \quad (21) $$

where $R$ is the reflectance at the surface of the etalon. FP etalons can be fabricated as thick films or thin crystals, on the end of a graded-index lens, or as a waveguide [Chen et al. 1990, Patela et al., 1995]. An etalon forms a cavity with a high $Q$, resulting in a transmission curve with a very sharp peak and steep sides. Values of $F \approx 100$ appear to be achievable for waveguide structures. We have used $F = 30$, which is easily attained, for the plot in Figure 5. The transfer curve is also very sensitive to the laser wavelength, and a laser source with a narrow linewidth is required for satisfactory operation. The bias point can be adjusted with either a voltage from a photovoltaic supply, or by tuning the laser wavelength. The tuning range for the laser should be large enough to tune across two or three adjacent transmission peaks. The wavelength spacing between adjacent peaks is called the free-spectral-range FSR, and is given by

$$ FSR = \frac{\lambda^2}{2nd}, \quad (22) $$
where $\lambda$ is the wavelength of the laser, and $n$ and $d$ are the index of refraction and the thickness or channel length for the etalon. The choice of the optimum bias point for a FP is more complicated than for the two-channel devices. The sensitivity, as given in Table 1, is proportional to $F^{1/2}G(F,S_{bias})$, where $G(F,S_{bias})$ is a numerical factor that depends on the bias point. It varies very slowly with $F$, for values of $F$ greater than 20. For the highest sensitivity, which occurs at $S_{bias}=1-S_{bias}^{R} \approx 0.75$, $G$ is 0.65. The superscripts refer to the transmitting and reflecting channels for the modulator. However, biasing the modulator at this point then leaves a nonzero value for the $S_{bias}^{T} - S_{bias}^{R}$ term in Equation (5), so that together with the steep slope of the transfer curve, the bias-control circuits would require a much higher bandwidth. This could lead to increased noise in the system.

If the modulator is biased at $S_{bias}^{T} = S_{bias}^{R} = 0.5$, in order to eliminate the difference term from Equation (5), $G$ is 0.50, and the sensitivity is reduced to about 77% of the maximum value.

The dynamic range is also affected by the choice in bias values. We calculated the dynamic range at both bias points in the same way as for the MZI modulator. For $S_{bias}^{T} = 0.5$, the harmonic at $2\omega$ is dominant, but when $S_{bias}^{T} = 0.75$, the harmonic at $3\omega$ dominates. For narrow-bandwidth applications, biasing the modulator at $S_{bias}^{T} = 0.5$ increases the dynamic range by about 3 dB, whereas biasing at $S_{bias}^{T} = 0.75$ increases the dynamic range by 10 dB for broadband applications. The best bias point will depend on the application, and should be determined experimentally, under actual measurement conditions.

### 4.4.5 Four-port coupler

Directional couplers also can be used as modulators for photonic EM-field sensors [Bulmer and Hiser, 1984], and are available as guided-wave devices in LiNbO$_3$. The four-port coupler is a common switching device for optical-fiber telecommunications. As normally fabricated, the passive bias point is at $S_{bias}^{X} = 1$, where the slope is zero and the coupler is insensitive to small signals. The superscript $X$ refers to the output for the cross-coupled state. Also, the coupling ratio is fairly insensitive to wavelength [Feuerstein et al., 1991], and the only practical mechanism for biasing this coupler is by applying a dc voltage from a photovoltaic cell. The transfer curve, shown in Figure 5 for the four-port coupler, has been shifted by adding an appropriate bias, so that its profile is more directly comparable to those for the other modulators. One way to fabricate a four-port coupler that is passively biased at $S_{bias}^{X} = 0.5$ is to shorten the coupling region. However, then the transfer curve is significantly changed [Bulmer and Hiser, 1984], so as to make it unsuitable for EM-field measurements.
The greatest sensitivity for the four-port coupler is obtained for $S^X_{\text{bias}} = 0.53$. This is close enough to the ideal bias at $S^X_{\text{bias}} = 0.5$ that the losses in sensitivity and dynamic range are insignificant, if the bias is chosen so as to eliminate the difference term in Equation (5). Although harmonics at both $2\omega$ and $3\omega$ are present, it is the component at $2\omega$ that limits the dynamic range, for broadband signals, to about 55 dB, as given in Table 1.

4.4.6 Three-port coupler

The transfer curve for the three-port ($1 \times 2$) coupler [Thaniyavarn, 1986] shows that it is well suited for EM-field measurements. The use of such couplers as field sensors has been reported [Howerton et al., 1988; Kanda and Masterson, 1992; Schwerdt et al., 1997]. These modulators do not require ancillary components, such as polarization splitters, Faraday rotators, or additional couplers. The bias point for best linearity and sensitivity ($S^A_{\text{bias}} = 0.5$) occurs at zero voltage, and is inherent in the design. Little data is available on the stability of the bias point with changes in temperature, and some form of active feedback control is expected to be necessary for critical field-measurement applications. The symmetry of the coupler makes its performance relatively insensitive to the laser wavelength, so bias control would be best implemented by using auxiliary photovoltaic cells. The coupler’s low sensitivity to the wavelength of the laser also means that there is a wide tolerance for the source selection or replacement. The sensitivity of the modulator, as given by the maximum slope of the transfer curve, is slightly better than for the MZI modulator. Our harmonic analysis shows the second harmonic, at $3\omega$, to be dominant, and as might be expected from the symmetry of the transfer curve, the harmonics at $2\omega$ and $4\omega$ are negligible. The calculated dynamic ranges for both wideband and single-frequency signals are comparable to those for the MZI modulator. Unfortunately, these couplers appear to be of limited interest to the telecommunications industry, and as far as we know, they are only available as special-order products.

5. Detectors and lasers

A high signal-to-noise ratio is a basic requirement for an accurate analog measurement system. The measurement systems we have discussed are biased to operate with power in the optical carrier, even in the absence of an RF signal. As seen in Section 2, high power in the optical carrier will increase the signal-to-noise ratio. PIN photodiodes, which can handle a few milliwatts of input power, are available. For frequencies up to 20 GHz, they would be the detector of choice for most applications. For higher-speed applications, Schottky-barrier photodiodes, with frequency responses up to 60 GHz, are commercially available, and frequency responses of 200 GHz [Li et al., 1993] and 500 GHz [Chou and Lin, 1992] have
been reported. The optical power available with present technology is high enough that the low-signal capabilities and higher noise of avalanche photodiodes usually are unwarranted.

Several types of lasers are available for use in EM-field sensor systems. As usual, the best choice requires a compromise between high performance and cost. If the bias is to be adjusted by controlling the laser wavelength, laser tunability is essential. The normalized, balance-detection signal processing reduces the noise due to laser-intensity fluctuations, but no system is completely effective, and a laser that also has a low RIN is still desirable.

5.1 External cavity-tunable diode lasers

Diode lasers, locked to a tunable external cavity, give excellent performance and have large tuning ranges. The output power obtainable is acceptable for most systems, and line widths of less than 100 kHz are achievable. Although rapid tuning over about 1 nm is usually available with piezoelectric transducers, tuning over a 50 nm range, with a mechanical stage, is much slower. The long-term durability of a mechanical-tuning mechanism is a question for a practical EM-field measurement system. These lasers are fairly expensive.

5.2 Distributed-feedback lasers

Distributed-feedback (DFB) lasers are a logical choice for many applications where a tuning range of only a few nanometers is adequate. The wavelength of a DFB laser can be controlled by varying its temperature with a thermo-electric cooler (TEC). We have tuned a DFB laser with 4 mW (6 dBm) of optical output at 1312 nm over 5 nm by this method. The frequency response of the feedback loop is limited by the thermo-electric-cooler response to about 1 kHz in this system. Faster feedback control can be obtained by varying the drive current to the laser, but this is limited in useful scan range to about 0.1 nm by the accompanying large fluctuations in laser intensity. There is some evidence that rapid fluctuations in the temperature, associated with the feedback control, are causing a gradual stress-related degradation in the output of our laser. The linewidths from DFB lasers depend on the quality of the Bragg grating formed during fabrication. Line widths of less than 50 MHz are commonly obtained, and are adequate except for a Fabry-Perot etalon with a high $Q$. DFB lasers can exhibit high RIN, unless an optical isolator is used to block reflections back from the fiber system.

5.4 Diode-pumped lasers

Optical losses from the modulator and ancillary components can easily add to more than 10 dB. Diode-pumped solid-state lasers can produce more than 50 mW
(17 dBm) out of a fiber, with very low RIN (≈150 dB), and provide excellent performance for applications that don’t require wavelength tuning. They are comparable in cost to the tunable, external-cavity lasers.

5.5 Fabry-Perot diode laser

For less-demanding measurements, a simple and considerably less expensive Fabry-Perot cavity laser, together with balanced detection, may provide an adequate signal-to-noise ratio [Schwerdt et al., 1997]. These lasers typically have multiple line outputs and higher RIN. Suitable performance may be achieved by selecting a laser having very few modes, and by operating it considerably above threshold, together with a modulator that is relatively insensitive to wavelength.

6. Conclusions

We have described system requirements and components suitable for optically linked, broadband EM-field sensors. Electrically short or resistively tapered dipole elements combine with the capacitive load impedance of an EO modulator to give a flat frequency response, and to preserve the phase and amplitude of the electric field with high fidelity. A number of photonic field-measurement systems have been reported in the literature, some of which use resistively tapered dipole elements. We summarize the characteristics of several systems in Table 2. The table is by no means exhaustive of the literature, but represents a spectrum of applications and performance. The systems mainly demonstrate concepts and feasibility, and are not necessarily engineered for best performance or stability. Because of this, detailed calibration-test results are not yet available. We have described how the performance

<table>
<thead>
<tr>
<th>Modulator</th>
<th>V_n (V)</th>
<th>Wavelength (nm)</th>
<th>P_d, power onto detector (mW)</th>
<th>Antenna</th>
<th>Bandwidth GHz</th>
<th>NEF * (V/m)</th>
<th>Reference</th>
</tr>
</thead>
<tbody>
<tr>
<td>Pockels Cell Bulk, Bi_4Ge_3O_12</td>
<td>2100</td>
<td>1300</td>
<td>1</td>
<td>RTD*</td>
<td>2</td>
<td>1</td>
<td>Kanda and Masterson [1992]</td>
</tr>
<tr>
<td>3-Port Coupler OGW*, LiNbO_3</td>
<td>10</td>
<td>850</td>
<td>0.006</td>
<td>RTD</td>
<td>4</td>
<td>0.1</td>
<td>ibid.</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>2 cm</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>Mach-Zehnder OGW, LiNbO_3</td>
<td>4</td>
<td>1300</td>
<td>1.5</td>
<td>metal rods</td>
<td>0.3</td>
<td>10^4</td>
<td>Kuwabara et al. [1992]</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td></td>
<td>14 cm</td>
<td></td>
<td></td>
<td></td>
</tr>
<tr>
<td>3-Port Coupler OGW, LiNbO_3</td>
<td>350 V/m</td>
<td>1300</td>
<td>0.08</td>
<td>segmented</td>
<td>3.5</td>
<td>3 x 10^4</td>
<td>Schwerdt et al. [1997]</td>
</tr>
</tbody>
</table>

*NEF: noise-equivalent field; *RTD: resistively tapered dipole;
OGW: optical guided-wave device
can be optimized by carefully choosing components, especially modulators with complementary outputs, and by using normalized, balanced-detection signal processing, together with active bias control. Detailed calibration-test results on one optimized system [Masterson et al., 1996a] will be available later. While dynamic ranges of about 100 dB have been achieved for some actual narrow-band systems, our analysis shows that it may be possible to extend that by another 15 dB for a fully optimized system. The wide bandwidths available with the photonic systems make them ideally suited for time-domain measurements of pulsed fields. They are equally suitable as reference sensors for antenna-calibration services at NIST, because they provide a purely capacitive load for the standard dipole, and the output can be filtered to remove interference from the environment of an open-area test site.

7. References


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