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FSK HETERODYNE SYSTEM EXPERIMENTS AT 1·5 μm USING A DFB LASER TRANSMITTER

Indexing terms: Optical communication, Optical transmission

140 and 70 Mbit/s FSK heterodyne system experiments are reported using single-filter detection with transmitter laser linewidths of 30 and 60 MHz. The system requirements that must be satisfied to avoid large performance penalties are considered.

Introduction: Earlier optical heterodyne system experiments, especially those reported by the British Telecom Research Laboratories,[1,2] have employed lasers with spectral linewidths much narrower than the bit rate. The line-narrowed semiconductor lasers used are, at present, relatively complex devices and this has led to considering the possibility of replacing them with DFB lasers. The advantage of the DFB laser is that it can be directly current-modulated but it has the disadvantage of having a relatively broad spectral linewidth. A heterodyne system experiment based on DFB lasers has recently been reported,[3] and this letter reports further work in this area. System results for experiments using IF linewidth to bit-rate ratios of 0·2 and 0·4 are presented, and the practical constraints imposed by envelope detection are discussed.

Experimental arrangement: This is shown in Fig. 1. The transmitter laser, a 1·5 μm ridge-waveguide distributed-feedback laser diode (DFB-LD),[4] was biased above threshold and the current was adjusted to give either a 30 MHz or a 60 MHz FWHM linewidth. FSK modulation of the DFB laser was achieved by directly modulating the bias current; a modulation current of 1 mA was used and this gave a FSK frequency deviation of 1·4 GHz. The output from the local oscillator laser, a 10 kHz linewidth tunable external cavity semiconductor laser (ext.cav.LD),[5] was combined with the output from the DFB laser in the single-mode coupler (SMC).[6] Owing to the sensitivity of the DFB laser to optical reflections more than 40 dB of optical isolation was needed at the output of this laser. To align the polarisation of the local oscillator with that of the DFB laser, a manual polarisation controller (pol.cont.) was inserted in the local-oscillator path. The output from the coupler was detected by a 560 MHz bandwidth PINFET receiver;[7] the local oscillator power incident on the photodiode was 20 μW. The highpass filter (HPF) PINFET receiver combination gave an IF bandwidth of 300 MHz. The envelope detector (env.det.) used for demodulating the IF consisted of a halfwave rectifier followed by a lowpass 100 or 50 MHz bandwidth filter depending on the bit rate in use. The baseband signal at the output of the IF demodulator was then regenerated and analysed by an error detector. The automatic frequency control (AFC) kept the IF constant at 420 MHz.

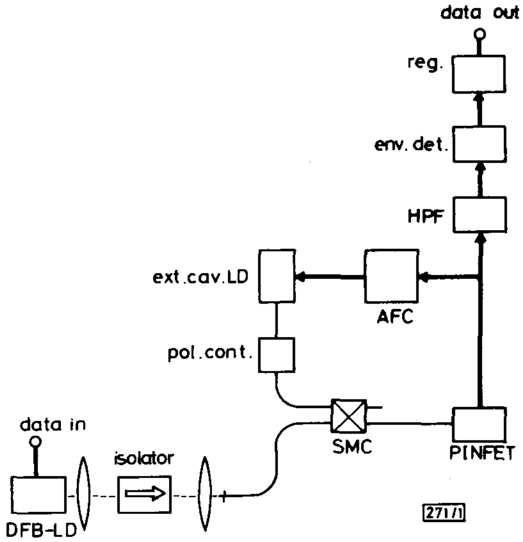

Fig. 1 *Experimental arrangement*

Discussion: If the receiver bandwidth is limited to the extent where it is necessary to use single-filter detection, and the FSK frequency deviation exceeds a certain critical value,[8] the receiver cannot distinguish between FSK or ASK modulation, and amplitude-demodulation techniques can be used at the IF. If the IF linewidth to bit-rate ratio exceeds 0·1% synchronous IF demodulation cannot be used and nonsynchronous techniques must be considered, of which envelope detection is the simplest. For the output from the envelope detector to be insensitive to both phase noise (IF linewidth) and IF value, the ratios of bit rate to IF, and IF linewidth to IF, should be sufficiently small. Another important consideration is the IF bandwidth. If this is too large the system performance will be degraded by the threshold effect of envelope detection; this is the point where the demodulation becomes nonlinear, but if the IF bandwidth is made too small the performance will be degraded by the IF linewidth. The nonlinear modulation characteristic of present DFB lasers causes the width of the modulation spectrum to be pattern-dependent; this effect can be treated as an effective broadening of the IF linewidth.

Results: Fig. 2 shows the measured bit error rate (BER) against mean received signal power for 140 and 70 Mbit/s modulation rates and different IF linewidths. At 140 Mbit/s the measured performances, at a BER of 10^{-8}, were -50 dBm and -46 dBm for 30 and 60 MHz linewidths, respectively. Comparing the shape of these two curves shows that the 60 MHz linewidth result has been degraded by the onset of error-rate saturation. The reason for this is not fully understood at present, but it is thought to be due to the optimum decision threshold having become signal dependent owing to the increased IF linewidth. The remaining BER curve is the 30 MHz IF linewidth, 70 Mbit/s result; the measured performance at a BER of 10^{-8} is -52 dBm. Owing to insufficient local-oscillator power the measured performances should be 2 dB worse than theory; the remaining 5 dB is attributed to the value of the bit rate to IF ratio. Owing to the values of this ratio, degradations associated with the IF linewidth to IF ratio have probably been larger than would normally be the case. The 70 and 140 Mbit/s results suggest that the envelope detection was approximately linear for the IF bandwidth to bit-rate ratios used because the measured results

DESIGN OF A WIDEBAND COMPACT SQUARE WAVEGUIDE POLARISER

Indexing terms: Microwave circuits and systems, Waveguide components

The computer-optimised design of a compact exponentially profiled E-plane depth corrugated square waveguide polariser is described which achieves $90° \pm 2°$ differential phase shift between the TE_{10}- and TE_{01}-wave for 11·8–12·5 GHz and 17·1–18·1 GHz. The input VSWR is better than 1·02 and 1·016 for the TE_{10}- and TE_{01}-wave incidences, respectively, between 10·3 and 18·8 GHz. The method of field expansion into suitable eigenmodes used considers the effect of higher-order mode interaction at all step discontinuities.

Introduction: Corrugated square waveguide polarisers[1-4] are extensively used for exciting circularly polarised waves in square aperture antennas. Owing to the periodic loading principle of the common structures,[1-4] good broadband impedance match is often difficult to meet. There is, however, increasing demand for polarisers capable of maintaining their properties over widely separated specific bands, such as the satellite bands.

In this letter, therefore, a polariser is suggested with a profiled corrugated depth structure—see Fig. 1. This type ensures good broadband impedance match. Moreover, the physical dimension of this specially profiled polariser is relatively compact. In contrast to the well established theories based on an anisotropic surface impedance of the corrugated geometry,[2,3,5,6] an accurate field theory method for direct computer-aided design is used.[4,7]

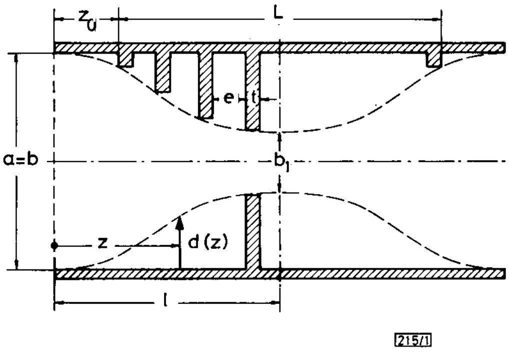

Fig. 1 *Square waveguide ($a = b$) polariser with continuously profiled depth $d(z)$ of the E-plane corrugated structure*

Theory: In order to appropriately describe the field in the individual waveguide sections, a number of higher-order modes excited at the discontinuities ($TE_{1n'}$, $TM_{1n''}$ and $TE_{0n'''}$, with $n' = 0, 2, 4, \ldots$, $n'' = 2, 4, 6, \ldots$, $n''' = 1, 3, 5, \ldots$) has to be taken into account. The calculated scattering coefficients show sufficient asymptotic behaviour if seven consecutive modes are considered.

At all discontinuities to be investigated, the field components are derived from the axial z-components of the magnetic and electric Hertzian vector potentials $\vec{\Pi}_h$ and $\vec{\Pi}_e$:[7]

$$\left.\begin{aligned} \vec{E} &= -j\omega\mu\nabla \times \vec{\Pi}_{hz} + \nabla \times \nabla \times \vec{\Pi}_{ez} \\ \vec{H} &= j\omega\varepsilon\nabla \times \vec{\Pi}_{ez} + \nabla \times \nabla \times \vec{\Pi}_{hz} \end{aligned}\right\} \tag{1}$$

where

$$\left.\begin{aligned} \Pi_{hz} &= \sum_{m=0}^{M}\sum_{n=0}^{N} A_{hmn} T_{hmn} + B_{hmn} T_{hmn} \\ \Pi_{ez} &= \sum_{m=1}^{M}\sum_{n=1}^{N} A_{emn} T_{emn} + B_{emn} T_{emn} \end{aligned}\right\} \tag{2}$$

For simplicity, in eqn. 2 the fields are written only for $z = 0$; the z-dependence in the forward and backward directions is understood.

The eigenfunctions T in eqn. 2 are normalised[7] so that the power carried by a given wave is proportional to the square of the mode coefficient A, B. By matching the tangential field components, given by eqn. 1, at the corresponding interfaces, the still unknown coefficients A, B in eqn. 2 may be determined after multiplication by the appropriate orthogonal function.[7] This yields the scattering matrix

$$(B) = (S)(A) \tag{3}$$

at the step discontinuity considered.

The scattering matrix of the total polariser structure is then calculated by directly combining the single scattering matrices.[7] This procedure preserves numerical accuracy, since this direct combination contains exponential functions with only negative argument.

Design and results: It is necessary to choose a profiled depth $d(z)$ of the E-plane corrugated structure so as to minimise the size of the polariser without significantly degrading the overall performance. To this end, we assumed here that $d(z)$ follows an exponential-type profile (cf. Fig. 1):

$$d(z) = \tfrac{1}{2}b\left\{1 - \exp\left[\left(\ln\frac{b_1}{b}\right)\left(\frac{z}{l} - \frac{1}{2\pi}\sin 2\pi\frac{z}{l}\right)\right]\right\} \tag{4}$$

The computer-aided design is carried out by an optimising program applying the evolution strategy method,[4] which varies the input parameters until the desired values of the differential phase shift $\Delta\phi$ and of the input VSWR, both for TE_{10} and TE_{01} incidence, respectively, for given midband frequencies and bandwidths, are obtained. Also, for a given number of irises, the parameters to be optimised are the waveguide housing dimensions $a = b$, the thickness t and distance e of the irises, as well as the depth profile parameter b_1, according to eqn. 4, and the position z_0 of the first iris.

A 21-iris polariser design for 90° differential phase shift at $f_1 = 12·1$ GHz and $f_2 = 17·7$ GHz is chosen for the design example. Fig. 2 shows the calculated differential phase shift and the VSWR curves for the TE_{10} and TE_{01} field incidences as a function of frequency. Between 11·8 and 12·5 GHz, and 17·1 and 18·1 GHz, respectively, the phase deviation is only $\pm 2°$ from the desired 90° value. The VSWR is better than 1·02 for TE_{10} incidence, and better than 1·016 for TE_{01} incidence, between 10·3 and 18·8 GHz. The overall length L of the polariser (cf. Fig. 1) is $L = 99·79$ mm.

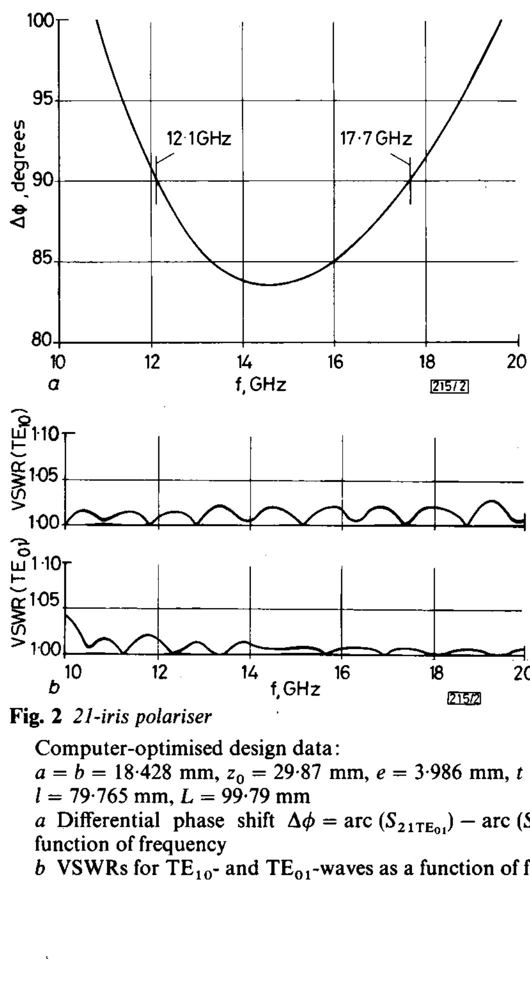

Fig. 2 *21-iris polariser*

Computer-optimised design data:
$a = b = 18·428$ mm, $z_0 = 29·87$ mm, $e = 3·986$ mm, $t = 0·955$ mm, $l = 79·765$ mm, $L = 99·79$ mm
a Differential phase shift $\Delta\phi = \text{arc}(S_{21TE_{01}}) - \text{arc}(S_{21TE_{10}})$ as a function of frequency
b VSWRs for TE_{10}- and TE_{01}-waves as a function of frequency