Measurements and modelling of free carrier lifetimes in Si and Si/poly-Si microrings

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SCATTERING FROM FINITE EXTENT FREQUENCY SELECTIVE SURFACES ILLUMINATED BY ARBITRARY SOURCES

Indexing terms: Antennas, Scattering

A method is proposed for the computation of the field scattered from a dichroic plate placed in the near-field region of the feeds. The primary pattern of the feed and dichroic plate assembly has been computed.

Introduction: Frequency selective surfaces, composed by arrays of metallic patches printed on dielectric films, may find interesting applications as diplexers in multifrequency reflector antennas, since they create spatially separated images of the focal region at different frequencies.[1] Generally these structures are characterised through their generalised scattering matrix, which implies that the surface is assumed to be infinite and with plane-wave illumination.[2] In practical antenna configurations, however, dichroic plates can be placed in the near-field region of the feeds and the incident wavefront cannot be considered plane.

Generally the problem is solved by computing the field in each point of the FSS (by invoking a principle of locality, the incident field is assumed to be a locally plane wave) and then by convolving the equivalent current distribution with the free-space Green function. However, since the behaviour of an FSS is related to a collective scattering phenomenon, the concept of locality does not seem to be applicable. Therefore an alternative procedure is proposed which is based on the use of the theorem of reciprocity.

With this formulation the finite extent of the FSS is automatically taken into account in the physical optics approximation; i.e. assuming that the induced current distribution on the patches is the same that one would have in the case of an infinite FSS.

Theory: In Fig. 1 the geometry of the problem is shown, where a finite extent planar FSS is illuminated by an arbitrary source a defined by an equivalent electric and magnetic current distribution $\boldsymbol{J}_a$, $\boldsymbol{M}_a$. In order to evaluate the scattered field, an electric dipole $\boldsymbol{J}_t$ (test source) is also introduced at the observation point $\boldsymbol{r}_t$ with orientation $\hat{\boldsymbol{t}}$. Now the Lorentz reciprocity theorem is applied to the volume V limited by the surface $S + S_i$, where only the incident field is considered for source a (i.e. the FSS is removed) and the total field for the test source (the FSS is present). After decomposing the field produced by the test source into an incident and a scattered component

$$(\boldsymbol{E}_t = \boldsymbol{E}_t^i + \boldsymbol{E}_t^s)$$

and

$$(\boldsymbol{H}_t = \boldsymbol{H}_t^i + \boldsymbol{H}_t^s)$$

one obtains the following result:

$$(\boldsymbol{E}_a - \boldsymbol{E}_a^i) \cdot \hat{\boldsymbol{t}} = \int_{S_r} \hat{\boldsymbol{z}} \cdot (\boldsymbol{E}_t^s \times \boldsymbol{H}_a^i - \boldsymbol{E}_a^i \times \boldsymbol{H}_t^s)\, dS - \int_{S_l} \hat{\boldsymbol{z}} \cdot (\boldsymbol{E}_t^s \times \boldsymbol{H}_a^i - \boldsymbol{E}_a^i \times \boldsymbol{H}_t^s)\, dS \qquad (1)$$

where S_l and S_r are the left and right faces of the FSS, respectively. This equation yields the effect produced by the FSS: the total field is obtained by adding the incident field due to the source a.

When the observation point is far from the FSS, the incident field $\boldsymbol{E}_t^i$, $\boldsymbol{H}_t^i$ can be approximated by a plane wave over the finite extent of the FSS, so that the scattered field $\boldsymbol{E}_t^s$, $\boldsymbol{H}_t^s$ can be readily computed by means of standard spectral techniques.

In fact, by extending the structure periodically, one can consider the problem of an infinite FSS illuminated by a plane wave coming from the observation direction. In this way the scattered field can be expanded in a series of Floquet modes, and eqn. 1 becomes

$$[\boldsymbol{E}_a(\boldsymbol{r}_t) - \boldsymbol{E}_a^i(\boldsymbol{r}_t)] \cdot \hat{\boldsymbol{t}} = A_r - A_l \qquad (2a)$$

where A_r and A_l are the surface integrals relative to the right and left face, respectively, and are given by

$$A_l = \sum_{pq} \begin{Bmatrix} V_{1pq} \\ V_{2pq} \end{Bmatrix} \begin{Bmatrix} \hat{\boldsymbol{K}}_{pq} \\ \hat{\boldsymbol{z}} \times \hat{\boldsymbol{K}}_{pq} \end{Bmatrix} \cdot \boldsymbol{F}_h(\boldsymbol{K}_{pq}) - \sum_{pq} \begin{Bmatrix} I_{1pq} \\ I_{2pq} \end{Bmatrix} \begin{Bmatrix} \hat{\boldsymbol{K}}_{pq} \times \hat{\boldsymbol{z}} \\ \hat{\boldsymbol{K}}_{pq} \end{Bmatrix} \cdot \boldsymbol{F}_e(\boldsymbol{K}_{pq}) \qquad (2b)$$

where

$$\boldsymbol{F}_e(\boldsymbol{K}_{pq}) = \int P(\boldsymbol{\rho}) \boldsymbol{E}_a^i(\boldsymbol{\rho}) \exp(-j\boldsymbol{k}_{pq} \cdot \boldsymbol{\rho})\, dS \qquad (3a)$$

$$\boldsymbol{F}_h(\boldsymbol{K}_{pq}) = \int P(\boldsymbol{\rho}) \boldsymbol{H}_a^i(\boldsymbol{\rho}) \exp(-j\boldsymbol{K}_{pq} \cdot \boldsymbol{\rho})\, dS \qquad (3b)$$

and $P(\boldsymbol{\rho})$ is the characteristic function of the FSS, which is 1 for points belonging to the FSS, and 0 otherwise.

The wavevector $\boldsymbol{K}_{pq}$ spans the points of the reciprocal lattice of the array. A_r has the same expression as A_l, but all the quantities are evaluated on the right face. In other words, the limited extent of the FSS is taken into account (in the physical optics approximation) by limiting the Fourier integrals (eqn. 3) to the actual FSS surface.

This approximation is widely used in the analysis of reflector antenna systems and is not acceptable only for the conducting patches close to the edge. However, the relative importance of these contributions should be negligible in practical cases where the FSS are very large in terms of the operating wavelength. The coefficients V_{mpq} and I_{mpq} are readily computed in terms of the generalised scattering matrix which provides a spectral characterisation of the FSS. With this formulation it is possible to analyse configurations where the FSS is placed in the near-field region of the feeds, where the evanescent fields are not negligible. However, it has been observed that the infinite double sum (eqn. 2b) in practical cases can be safely truncated to the first orders of evanescent Floquet modes.

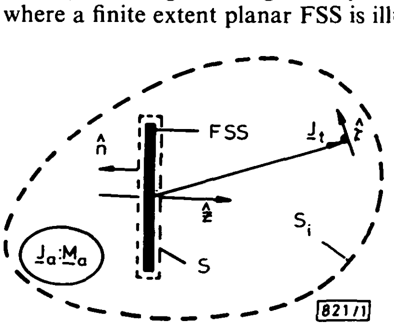

Fig. 1 *Geometry of the problem*

Source a is indicated by its equivalent currents $\boldsymbol{J}_a$ and $\boldsymbol{M}_a$; test source is indicated by $\boldsymbol{J}_t$

Results: The previous formulation has been applied to the arrangement shown in Fig. 2, which serves as a feeding system

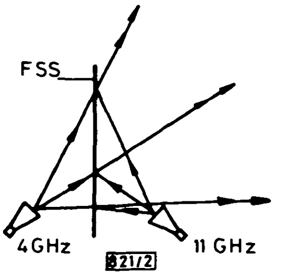

Fig. 2 *Geometry of the 4/11 GHz feed system analysed*

Feeds are placed 20 cm from the FSS and tilted by 35·17° with respect to FSS normal; *E*-field in offset plane

for an offset reflector antenna. The FSS is composed of an array of crossed dipoles of length $L = 11{\cdot}13$ mm and width $W = 0{\cdot}10$ mm arranged in a square lattice with size $d = 8{\cdot}79$ mm. The conductors are printed on a Kapton film (thickness 0·5 mm, $\varepsilon_r = 3{\cdot}2 - j0{\cdot}02$) supported by a Kevlar honeycomb structure (thickness = 6.35 mm, $\varepsilon_r = 1{\cdot}05$) protected by two Kevlar films (thickness 0·18 and 0·25 mm, $\varepsilon_r = 4{\cdot}1 - j0{\cdot}024$). The characteristics of the dielectric layers are fixed by technological constraints. With these dimensions, the FSS has a −0·5 dB transmission bandwidth extending up to 4·5 GHz and a −0·5 dB reflection bandwidth of 1·4 GHz centred at 11·2 GHz. The feeds are rectangular and placed at such a distance from the FSS that the wavefronts cannot be considered as planar.

Using the technique described before, the transmitted and reflected primary patterns have been computed, and an example is shown in Fig. 3. It can be noted that in the *E*-plane

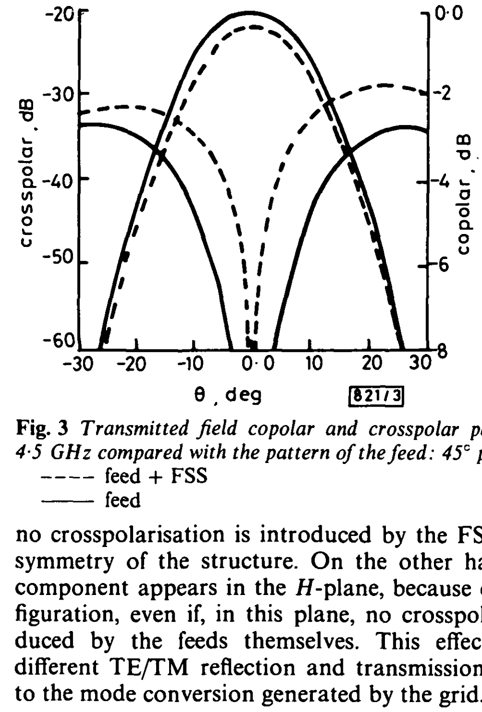

Fig. 3 *Transmitted field copolar and crosspolar pattern at frequency 4·5 GHz compared with the pattern of the feed: 45° plane cut*

---- feed + FSS
—— feed

no crosspolarisation is introduced by the FSS because of the symmetry of the structure. On the other hand a crosspolar component appears in the *H*-plane, because of the offset configuration, even if, in this plane, no crosspolarisation is produced by the feeds themselves. This effect is due to the different TE/TM reflection and transmission coefficients and to the mode conversion generated by the grid.

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EFFECT OF INTERMODULATION IN MULTICHANNEL OPTICAL HETERODYNE SYSTEMS

Indexing term: Optical communications

It is shown that large numbers of optical heterodyne channels may be frequency multiplexed over a single-mode optical fibre without the intermodulation byproducts of the photodetection process becoming a problem, provided the channels are spaced at five times their optical bandwidth, or greater.

Introduction: Frequency division multiplexing (FDM) techniques might at first sight appear to offer a very efficient means of utilising the vast transmission capacity of modern single-mode optical fibres. The generation of large numbers of closely spaced optical FDM channels now seems possible due to the progress that has been made recently in fibre optic heterodyne detection techniques.[1] Indeed, the use of closely spaced FDM channels is probably the only efficient way of exploiting a fibre's full capacity. Unfortunately, light wave receivers such as photodiodes respond to the incident instantaneous optical power and are therefore nonlinear (or square-law) devices. This square-law nonlinearity leads to the generation of intermodulation terms in the photoreceiver output spectrum if the incident optical signal comprises closely spaced FDM channels. This poses the question: what channel spacing allows maximum utilisation of the fibre's capacity, and, at the same time, minimises the parasitic intermodulation and crosstalk noise which can occur in these systems? It is the aim of this letter to address this question. The receiver model used in the following analyses is shown in Fig. 1.

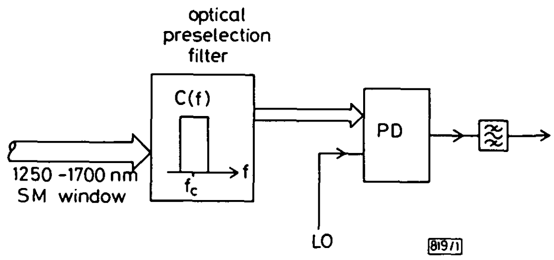

Fig. 1 *Receiver model*

PD is the photodetector and SM is the single-mode fibre

Theory: It is assumed that the incident optical fields are plane, spatially coherent waves which, during detection, are normal to the receiver area. Thus, only temporal effects of the received radiation need to be considered as the receiver operates as a point detector.[2] The photodetector bandwidth is assumed to be large enough to pass the wanted signal without distortion.

The photodetector output response $r(t)$ is proportional to the intensity or low-frequency part of the square of the modulus of the incident electric field vector $s(t)$,[3] i.e.

$$r(t) \propto L\{|s(t)|^2\} \tag{1}$$

where $L\{.\}$ means low-frequency part. (Quantum mechanical analysis shows that the rapidly varying sum, and double, frequency components never appear for the type of photodetector considered in this letter providing $h\nu \gg kT$.[3])

The optical signal spectrum incident on the photodetector is shown in Fig. 2.

The purpose of the optical bandpass filter shown in Fig. 1 is to coarse tune the receiver to the required part of the spectrum and reject most of the unwanted channels. This preselection process helps to reduce the power density falling on the photodetector and therefore, the total intermodulation and shot noise. (It is assumed here that the optical filter does not have sufficient selectivity to reject all unwanted channels.)

A suitable description of the optical bandpass filter in the frequency domain is

$$C(f) = \delta(f \pm f_c) * E(f) = E(f \pm f_c) \tag{2}$$

where $*$ denotes convolution, and $E(f)$ is the spectral shape of