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F. Delorme and S. Slempkes (*Centre National d'Etudes des Télécommunications, 196 Avenue H. Ravera, BP 107, 92225 Bagneux Cedex, France*)

P. Gambini and M. Puleo (*Centro Studi e Laboratori Telecommunicazioni, Via G. Reiss Romoli 274, 10148 Torino, Italy*)

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INFLUENCE OF OPTICAL AMPLIFIER LOCATION ON FOUR-WAVE MIXING INDUCED CROSSTALK

C. Bungarzeanu

Indexing terms: Nonlinear optics, Optical amplifiers, Optical transmission, Optical communication

Following the optical amplifier position in the transmission link, the four-wave mixing (FWM) induced crosstalk is increased up to two times the optical gain. A simple approximate expression of the crosstalk penalty is derived and compared with exact results for standard and dispersion shifted (DS) optical fibres.

Introduction: Optical frequency division multiplexing (OFDM) techniques combined with optical amplifiers acting as multiwavelength all-optical repeaters, are very promising for fibre optic communication systems.

However, nonlinear interactions in the transmission fibre lead to fundamental limitations on system performance. Theoretical approaches [1, 2] and several experiments reported in the literature [2, 3] have identified FWM as the dominant nonlinear process for many OFDM systems. Thus, the limitation on transmitter power due to the FWM interaction is determined by the FWM induced crosstalk between signal channels.

On the other hand, the presence of an optical amplifier in the transmission link increases the FWM crosstalk, unless the amplifier is located at the end of the link. As the position of the optical amplifier depends on field deployment conditions, cost-effectiveness, upgrading scenarios and other considerations, the resulting FWM crosstalk penalty should be taken into account, to modify, if necessary, the system configuration.

The purpose of this Letter is to analyse the dependence of FWM crosstalk penalty on the amplifier gain and location, system parameters and fibre characteristics.

Theory: Assuming the same input power P, for each OFDM channel, the FWM generated power in the middle channel, at the end of x km of fibre can be written as [1]

$$P_{fwm}(x) = K_{fwm}(x)10^{-\alpha x/10}P^3 \tag{1}$$

where K_{fwm} express the FWM interaction and depends on the total number of channels, the frequency spacing, and fibre characteristics such as the attenuation coefficient α and the chromatic dispersion.

The system illustrated in Fig. 1 includes an optical amplifier of gain G, situated at distance x from the transmitter in a link

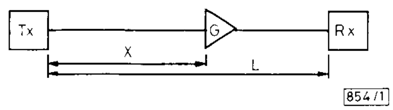

Fig. 1 *System configuration*

of total length L. In this case, the FWM generated power at the end of the link is

$$P'_{fwm}(x) = P^3 10^{(G-\alpha L)/10}\{K_{fwm}(x) + K_{fwm}(L-x) \times [1 + K_{fwm}(x)P^2 10^{-\alpha(L-x)/10}]^3 10^{(2G-2\alpha x)/10}\} \tag{2}$$

and the FWM induced crosstalk

$$X(x) = \frac{P'_{fwm}(x)}{10^{(G-\alpha L)/10}P} \tag{3}$$

The amplifier induced crosstalk penalty in decibels is defined as

$$\Delta X(x) = 10 \log \frac{X(x)}{\min\limits_{x} X(x)} \tag{4}$$

If K_{fwm} is assumed constant, and neglecting the term that contains P^2 in eqn. 2, the crosstalk penalty can be expressed by the following approximation:

$$\Delta X(x) \simeq 10 \log (1 + 10^{2(G-\alpha x)/10}) \tag{5}$$

Results: To calculate the FWM crosstalk penalty due to the optical amplifier, the following OFDM system has been considered: eight channels centred on 1550 nm with 10 GHz frequency spacing. The total length of the link is $L = 100$ km and the fibre attenuation coefficient is $\alpha = 0{\cdot}2$ dB/km. According to the fibre type, the chromatic dispersion C takes two values: 15 ps/nm km for standard fibre and 3 ps/nm km for DS fibre. The computation of K_{fwm} is based on the model proposed in Reference 1.

The results, corresponding to a DS fibre are shown in Fig. 2, with dotted lines for four different input power values ($P = 5$, 10, 12·5, and 15 dBm, respectively). For comparison, the approximation (eqn. 5) is plotted with a solid line. It is apparent in this case than for input powers below 5 dBm the error between the exact calculation and the approximation is less than 1 dB.

Fig. 3 shows the crosstalk penalty behaviour for different optical gains at 5 dBm input power. Exact calculations were performed for standard and DS fibres. Again, the approximation (eqn. 5) is plotted for comparison. As can be seen, the maximum error between the exact calculations and the approximation is less than 3 dB if the original gain is less than αL.

Conclusions: For input powers per channel of practical interest (below 5 dBm) and for optical gains less than αL, the amplifier induced crosstalk penalty is described by the approximation of eqn. 5 with more than 3 dB accuracy. This means that for x greater than G/α the amplifier induced crosstalk penalty becomes negligible. The conclusion is that if more liberty is needed in choosing the amplifier location, then the

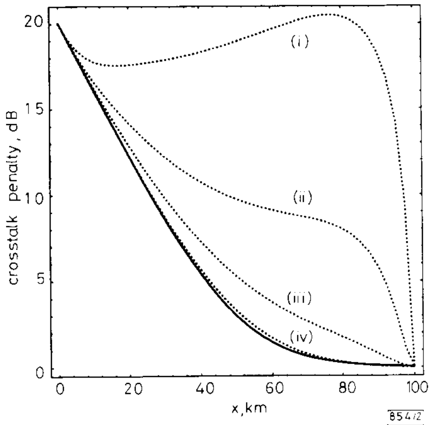

Fig. 2 *FWM crosstalk penalty for $G = 10\,dB$, 8 channel OFDM system, $L = 100\,km$, $\alpha = 0{\cdot}2\,dB/km$, $C = 3\,ps/nm\,km$*

······· exact calculation
(i) $P = 15\,dBm$
(ii) $P = 12{\cdot}5\,dBm$
(iii) $P = 10\,dBm$
(iv) $P = 5\,dBm$
——— approximation (eqn. 5)

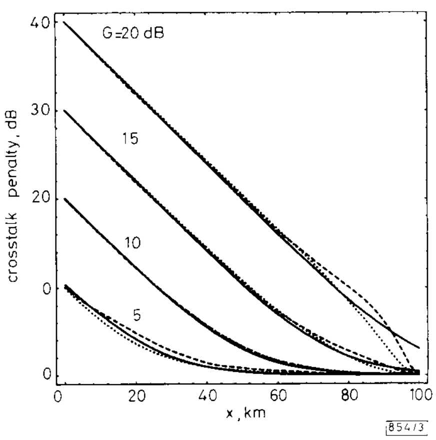

Fig. 3 *FWM crosstalk penalty at 5 dBm input power and different optical gains*

------- $C = 15\,ps/nm\,km$ (standard fibre)
······· $C = 3\,ps/nm\,km$ (DS fibre)
——— approximation (eqn. 5)

optical gain should be reduced accordingly. Thus, whereas the amplifier needs to be located at a distance d from the transmitter, a crosstalk penalty less than 3 dB is assured only for an optical gain less than αd.

Moreover, these results can be easily applied to OFDM links with more than one optical amplifier to determine the overall crosstalk for planning and design purposes.

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C. Bungarzeanu (*Ecole Polytechnique Fédérale de Lausanne, DE-TCOM, 1015 Lausanne, Switzerland*)

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BROADBAND UNIPLANAR HYBRID RING COUPLER

C.-H. Ho, L. Fan and K. Chang

Indexing terms: Waveguide couplers, Directional couplers

A new uniplanar hybrid ring coupler using coplanar waveguide (CPW) and slotline is presented. This new coupler provides substantially improved amplitude and phase characteristics over a broad bandwidth compared with the conventional microstrip hybrid ring couplers. Experimental results show that the new coupler has a bandwidth of approximately one octave from 2 to 4 GHz with $\pm 0{\cdot}25$ dB power dividing balance and $\pm 1°$ phase balance.

Introduction: The microstrip rat-race hybrid ring coupler is the basic power divider in printed microwave integrated circuits. The 20–25% bandwidth of this coupler limits its applications to narrowband circuits. Several design techniques have been developed to extend that bandwidth. One technique used a $1/4\lambda_g$ coupled microstrip line section to replace the $3/4\lambda_g$ section of the conventional $3/2\lambda_g$ microstrip ring coupler, where λ_g is the guide wavelength [1]. Although the bandwidth was increased to approximately an octave, the difficulty of constructing the coupled microstrip line section, which required short circuits at the ends, limited its use to lower frequencies. Another modified version of the microstrip rat-race hybrid ring substituted a quarter-wavelength slotline section etched on the other side of the substrate for the phase delay section [2]. Although a bandwidth of two octaves was realised, the two-sided substrate required a more complicated photolithographic process and limited the coupler to lower frequencies. Two other approaches [3, 4] used hypothetical ports with matching circuits. This technique achieved a 50% bandwidth, however the matching circuits described in Reference 3 required very wide microstrip lines and a large number of different impedances, and the broadband design technique presented in Reference 4 was useful in the sum mode of operation only. Both matching techniques also demanded intensive optimisation to obtain good performance.

To extend the bandwidth with a simple design procedure and uniplanar structure, this Letter presents a new uniplanar hybrid ring coupler consisting of a slotline ring with one slotline feed and three CPW feeds. The design technique substitutes one reverse-phase slotline-T junction for the conventional rat-race phase delay section. Because the phase reverse of the slotline-T junction is frequency independent, the resulting slotline ring coupler has a broad bandwidth. Experimental results presented in this Letter verify the concept and show a usable bandwidth of more than one octave.

Circuit design and measurements: Fig. 1 shows the physical configuration of the uniplanar hybrid ring coupler. As shown in Fig. 1, the E-arm of the uniplanar slotline ring is fed through a CPW line connected with one CPW-to-slotline transition. The slotline T junction shown in Fig. 1 is used as a phase inverter. The impedance of the slotline ring is given by

$$Z_s = \sqrt{(2)}\,.\,Z_{CPW} \tag{1}$$