Optical Demultiplexing for Subcarrier Multiplexed Systems

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***Abstract*— Subcarrier multiplexing (SCM) is an attractively simple technique for establishing multiple independent channels over a single fiber. At the receiver, electrical mixing converts a selected channel to baseband. The photodiode, oscillator, mixer, and preamplifier must, however, operate up to the highest subcarrier frequency. The use of an optical filter for channel selection allows any subcarrier to be accessed with only baseband electronics. To obtain the best performance with this approach, sub-subcarrier modulation is needed. In this paper we propose a method suitable for externally modulated CATV systems and show that the sensitivity is comparable to the best conventional receivers.**

I. Introduction

IN conventional SCM systems, every baseband signal is translated to a unique subcarrier frequency by allowing each to modulate a different microwave subcarrier. The subcarriers are then summed and the composite signal is used to modulate an optical source, either by directly modulating the injection current of a semiconductor laser or by means of an external modulator. At the receiver a photodiode with a bandwidth greater than the highest subcarrier frequency is used to directly detect the composite signal intensity. An individual channel may then be selected by electrically mixing with a microwave oscillator and filtering. The photodiode and any preamplifier need adequate bandwidth to pass all the subcarrier frequencies. For multichannel CATV systems this may be several gigahertz. In a previous publication [1] we reported the experimental demonstration of an alternative method, optical prefiltering, in which a narrowband optical filter such as a Fabry-Perot filter (FPF) is used to demultiplex the desired channel. Modulation of an optical source by a microwave subcarrier generates sidebands in the optical spectrum, centerd on the optical carrier and separated from it by the subcarrier frequency. Modulation of the subcarrier produces subsidiary sidebands around the primary sideband. If an optical filter is used to select a particular subcarrier's sideband and its associated modulation sidebands, as shown in Fig. 1, and this signal is incident on a photodiode, the photocurrent will contain the original baseband information. Since the filter blocks all the other optical sidebands the photodiode is effectively illuminated by a single modulated signal. Detection occurs at baseband and the photodiode and electronics need only have the channel bandwidth. Channels are selected by tuning the optical filter. This scheme removes the need for good receiver intermodulation performance since only one channel is detected.

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Although optically prefiltered receivers can be used in place of conventional receivers without other system modifications, as described above and previously [1], the performance in this case is not ideal. Section II of this paper considers the optimum external modulator bias for prefiltered systems and shows that this differs from the conventional case. To transmit high quality analogue signals efficiently in an SCM system, frequency modulation (FM) must be used since much lower subcarrier modulation depths are needed to give acceptable intermodulation levels with amplitude modulation (AM). However, although straightforward FM can be used in an optically prefiltered system, it does not give an increase in signal to noise ratio (SNR) over carrier to noise ratio (CNR) and so Section III looks at sub-subcarrier techniques to allow this to be achieved. In Section IV, the sensitivity of an optically prefiltered receiver in a system using the proposed modulation system is compared with that of a conventional SCM receiver for FM TV. Finally, Section V addresses the channel spacing that can be achieved with FPF's.

II. Optimum External Modulator Bias

Typically, an external modulator consists of an interferometer implemented in integrated optic form in an electro-optically active material. An applied voltage produces equal but opposite electric fields across the two arms of the interferometer, advancing optical phase in one and retarding it in the other. Fig. 2 shows the amplitude and intensity transfer characteristics for a modulator plotted against the phase difference between the arms. Since direct detection systems respond to optical intensity, it is normal to consider the device as a (nonlinear) intensity modulator. It is biased at the -3 dB point marked on Fig. 2 where the gradient is most linear. The antisymmetry of the curve about this point ensures that there are no even intermodulation products in the detected signal. If the optical spectrum is considered, it is seen there are components on either side of the optical carrier at multiples of the modulation frequency but that these mix at the photodiode and cancel to give a linear response. Optical filtering radically alters this by rejecting harmonic terms and the nonlinearity now causes intermodulation. At a modulator bias, θ, (x-axis of Fig. 2), the spectrum produced in the field, E, of an optical carrier of frequency, ω_o, by modulation with

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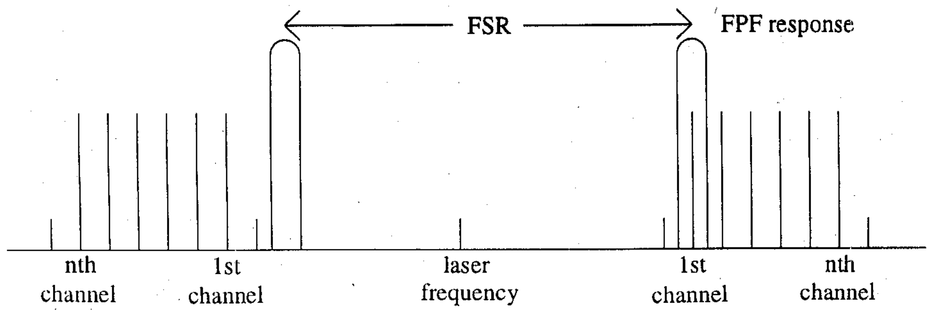

Fig. 1. Selection of 1st of n SCM channels in optical spectrum using FPF.

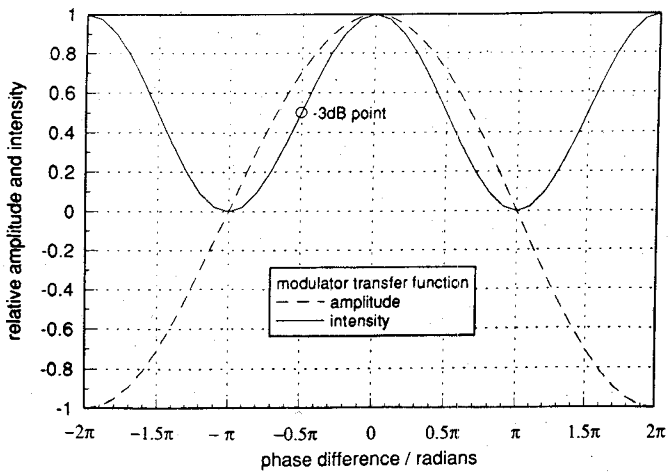

Fig. 2. External modulator characterstics.

frequency, ω_m, and depth, m_E, takes the form

$$E = \cos\left(\frac{\theta}{2}\right) J_0(m_E)\cos(\omega_o t) \pm \sin\left(\frac{\theta}{2}\right) J_1(m_E)\cos(\omega_o t \pm \omega_m t) + \cos\left(\frac{\theta}{2}\right) J_2(m_E)\cos(\omega_o t \pm 2\omega_m t) \pm \cdots. \quad (1)$$

It can be seen that making $\theta = \pm\pi$ radians will maximize the desired $\omega_o + \omega_m$ component and remove all even terms. This is double sideband suppressed carrier (DSBSC) modulation and no power is transmitted at the optical carrier frequency. The conventional bias point of $\theta = -\pi/2$ radians results in 3 dB less optical power at the sideband frequencies. When biased for DSBSC the behavior is similar to an unprefiltered system, with no even order intermodulation and the modulation depth is limited by the acceptable 3rd order intermodulation level.

III. Sub-Subcarrier Modulation

Since, in a prefiltered receiver, the optical filter passes only the selected optical sideband and not the carrier there is no mixing between them and modulating the subcarrier frequency does not produce an FM signal at the photodetector. An optically prefiltered receiver can be used to directly demodulate FM signals by slope detection. One side of the filter passband is aligned with the sideband center frequency. This converts the FM signal to AM, but without amplitude limiting the improvement between CNR and demodulated SNR possible with FM is not realized. Restricting the frequency deviation to the most linear portion of the filter response results in lower modulation depth of the received signal than for AM and significant insertion loss. A further drawback to slope detection is that operating on the side of the filter characteristic converts laser phase noise to intensity noise.

In order to realize the noise advantage of FM in an optically prefiltered SCM system it is necessary to use a sub-subcarrier modulation technique. The signal to be transmitted modulates the frequency of the sub-subcarrier. At the receiver, the amplitude modulated sub-subcarrier is detected and the resultant signal is applied to an FM demodulator, exactly as for a non-prefiltered receiver. Thus no additional electronic processing is needed. Three modulation techniques are considered. We assume the use of an Nd:YAG laser with an external modulator biased for DSBSC since this offers the best performance.

An obvious sub-subcarrier scheme is to allow each baseband FM channel to amplitude modulate an RF subcarrier. The optical filter tuned to the desired subcarrier channel, i.e. its center frequency is offset from the laser frequency by the subcarrier frequency and the frequency modulated upper and lower sidebands of the subcarrier sideband mix with it at the photodetector to regenerate the FM baseband signal. This has two potential drawbacks. First, the transmission of upper and lower subcarrier sidebands means that the optical bandwidth and, therefore, the channel spacing is twice as great as is necessary, halving the number of channels which can be transmitted with a given external modulator bandwidth. Second, since true AM is used and the photodiode acts as a perfect square-law detector, significant levels of second harmonic distortion are generated by mixing between the upper and lower sidebands. 100% sub-subcarrier modulation is desirable to maximize the received signal. For AM, this results in a second harmonic distortion level only 12 dB below the fundamental.

A variation on this technique is to use balanced mixers to generate DSBSC subcarriers. No power is transmitted at the optical sideband frequency, and mixing at the photodiode is between the upper and lower subcarrier sidebands only. In this system the recovered baseband signal is undistorted (it is entirely second harmonic) but the mean frequency and frequency deviation have been doubled. Thus either the frequency modulators or the demodulators must be nonstandard. The optical bandwidth of the channel is the same as for AM

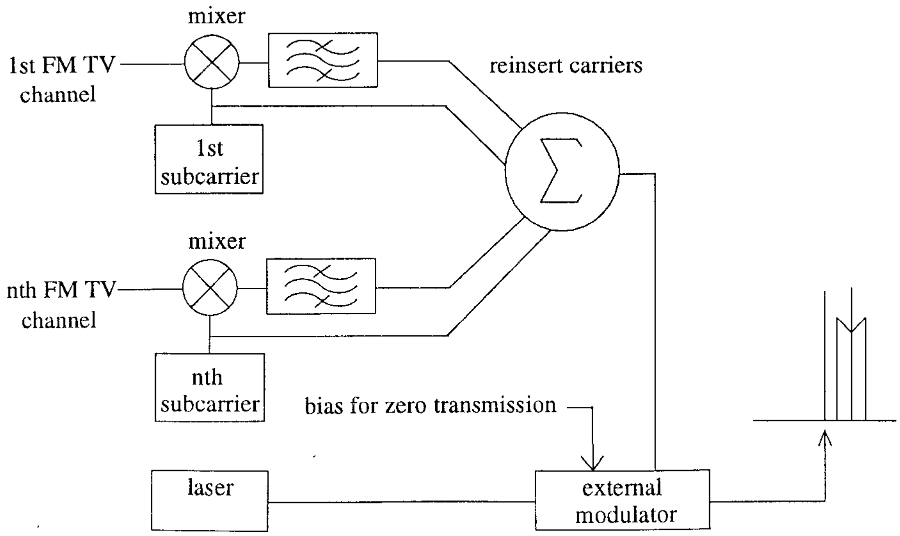

Fig. 3. Sub-subcarrier SCM FM TV transmitter.

and, assuming the same peak optical power per channel, the detected signal level is also the same.

A better approach is to directly modulate the frequency, ω_m, of the subcarrier, and to insert an unmodulated carrier at a slightly offset frequency, ω_u, as shown in Fig. 3. The optical filter is tuned to pass one modulated optical sideband and the associated unmodulated sideband. These mix to produce an FM output signal with a center frequency equal to their frequency difference. If the subcarriers have modulation depths, m_E, then the modulated field (retaining only the first Bessel function terms) is

$$E = A \cdot 2J_1(m_E) \cdot (m_E) \cdot (\cos \omega_{sc} t + \cos \omega_{ro} t) \cos \omega_o t. \quad (2)$$

For the largest individual channel modulation indices which can be used without unacceptable distortion (Section IV), making the approximation $2J_1(x) \cong x$ introduces an error in the calculated detected power of less than 2%. The field comprises an upper and lower optical sideband, only one of which is passed by the optical prefilter. Denoting the optical frequency of the unmodulated sideband as ω_{rx} and that of the modulated sideband as $(\omega_{rx} + \omega_{\text{sig}})$ then the modulated optical field, E_{rr}, passed by the prefilter becomes

$$E_{rx} = \frac{Am_E}{2}[\cos \omega_{rx} t + \cos(\omega_{rx} t + \omega_{\text{sig}} t)]. \quad (3)$$

The signal current produced in a photodiode of responsivity, R, is

$$i_{\text{sig}} = \frac{R(Am_E)^2}{4} \cos \omega_{\text{sig}} t. \quad (4)$$

The received power is the same as for AM and DSBSC but there is no harmonic distortion and no significant bandwidth increase is entailed.

IV. Comparison of Sensitivity

In order to compare the sensitivity of an optically prefiltered receiver with that of a conventional one, two aspects must be considered; the receiver noise levels and the signal photocurrent produced at the receiver by a given channel power. To find the signal current for each receiver it is necessary to determine the maximum acceptable modulation index. This is system dependent; we will consider high quality FM video with a SNR of 56 dB. To achieve this a CNR of 17.5 dB is needed for a 30-MHz signal bandwidth [2].

A. Conventional SCM Receiver

When N independent channels, each with an intensity modulation index of m_I, are combined an rms modulation depth, μ, may be defined as

$$\mu = \sqrt{\frac{N}{2}} \cdot m_I. \quad (5)$$

For a mean optical power at the receiver of P_o, the optical signal is then

$$P = P_o + P_o m_I \sum_{p=1}^{N} \cos \omega_p t \quad (6)$$

and the mean square signal current for an individual channel is

$$\langle i_{\text{sig}}^2 \rangle = \frac{(RP_o m_I)^2}{2}. \quad (7)$$

Alameh and Minasian [3] have reported a detailed statistical analysis of nonlinear distortion as a function of μ for external modulators. Their results are presented in the form of carrier to nonlinear distortion ratios (CNDR). At the relatively high modulation indices which can be used with FM systems, they find little difference between linearised and nonlinearised external modulators. To allow for noise, the CNDR must be greater than 17.5 dB. For FM systems, such as are considered here, the modulation indices are sufficiently high that the shot noise power (which is determined by the mean photocurrent for all channels) is much less than the signal power even for quite large numbers of channels. Laser relative intensity noise

(RIN) may also be neglected with an externally modulated Nd:YAG source. Thus the receiver is the dominant noise source and, writing the mean square noise current as $\langle i_n^2 \rangle$, the overall CNR is given by

$$\frac{1}{\text{CNR}} = \frac{2\langle i_n^2 \rangle}{(RP_o m_I)^2} + \frac{1}{\text{CNDR}(\mu)}. \quad (8)$$

The rms modulation depth can be optimized for the best system sensitivity (independent of the number of channels and the receiver noise level) by substituting μ for m_I in (8) using (5), imposing the condition that CNR =17.5 dB, and then finding the value of μ which minimizes the required ratio of P_o^2 to $\langle i_n^2 \rangle$. This minimum occurs for a CNDR of 22 dB [3] and a value for μ of about 0.55. In the case of a 60 channel system the corresponding value of m_I is 0.1, while for 10 channels it is 0.24.

There are two possibilities for the receiver. Avalanche photodiodes (APD's) are inappropriate for SCM receivers because of the very wide bandwidth required, the high linearity needed to avoid intermodulation, and because the otherwise negligible shot noise is increased, relative to the signal, by the excess noise factor. Good noise performance can be achieved with a nonresonant PINFET receiver. An effective load resistance of 300 ohms is suggested [4] as a maximum value for a SCM system in order to maintain sufficient dynamic range and linearity. This is the dominant noise source for reasonably low capacitance devices [5]; a typical value for the total mean square noise current spectral density is about 60 (pA)^2/Hz for a good device capable of operation to several GHz. Better performance is offered by resonant PINFET structures [5], [6]. These utilize a small inductor either bonded in parallel with the photodiode or in series between the photodiode and the FET gate. Parallel connection allows shot noise limited operation at the resonant frequency but the bandwidth is very restricted. Series connection can maintain better overall performance over an octave bandwidth giving about 5 dB improvement over the nonresonant PINFET [4], [5]. The main disadvantage is that device fabrication requires a small (≅nH) inductor to be bonded to the photodiode and FET and tuned to match the device capacitances. Dynamic range also remains a consideration. For comparison with an optically prefiltered receiver, then, we consider receivers with mean square noise current spectral densities of 60 (pA)^2/Hz (nonresonant PINFET) and 20 (pA)^2/Hz (resonant PINFET). A total CNR of 17.5 dB with a CNDR contribution of 22 dB requires the receiver noise to be about 19.5 dB below the carrier power. Thus, for a 30-MHz bandwidth and taking a typical responsivity value for an InGaAs detector at 1500 nm of 0.9 A W^{-1}, the minimum mean optical powers required, for 60 channels and 10 channels respectively, are about −22 dBm and −26 dBm in the case of a nonresonant receiver and −28 dBm and −32 dBm for a resonant receiver.

B. Optically Prefiltered Receiver

The amplitude characteristic of an external modulator takes the same sinusoidal form as its intensity characteristic (Fig. 2) and the harmonic distortion and intermodulation produced in the optical field by a modulation index, m_E, is the same as that in the optical intensity for an equal value of m_I (but requires 6 dB more electrical drive power). Thus, the same intermodulation analysis [3] may be applied except that we now choose to define the rms modulation depth as

$$\mu = \sqrt{N} \cdot m_E. \quad (9)$$

The factor of 2 has been omitted to keep the definition of N, the number of modulated channels, consistent; for every channel there is also an unmodulated carrier which also contributes to the rms modulation depth. To compare receiver sensitivities it is necessary to relate the field amplitude, A, in (4) to the mean optical power, P_o, used in the analysis of the conventional receiver. Since the modulator is conventionally biased at the −3 dB point, P_o is half the maximum power that could be delivered to the receiver. Maximum power is equal to the square of the rms field amplitude and so the mean square signal power is

$$\langle i_{\text{sig}}^2 \rangle = \frac{(RP_o)^2 m_E^4}{2}. \quad (10)$$

Since the signal power increases more rapidly with the modulation index than is the case for the conventional receiver, the rms modulation depth for the best system sensitivity is shifted very slightly in the direction of greater modulation; from Alameh's and Minasian's results [3], the optimum value of μ is 0.6. For a 60 channel system m_E is just under 0.08, while for 10 channels it is about 0.19. Comparing (10) with (7) it can be seen that there is a significant sensitivity penalty for the prefiltered system, since the smaller modulation index is raised to the fourth power rather than the second. The penalty grows with the number of channels since m_E must be reduced to maintain constant μ; 8 dB more optical power is needed for 10 channels and 12 dB more for 60 channels. This is offset, however, by the considerable improvement in receiver sensitivity which can is possible.

For an optically prefiltered system, the best performance is offered by an APD receiver. A large bandwidth is not required, linearity is much less critical since the optical filter blocks all but the selected channel, and only the shot noise of that channel is multiplied by the excess noise factor. In order to estimate the sensitivity of a prefiltered receiver we consider only what can be achieved using currently available commercial components. A typical InGaAs APD is the Mitsubishi PD805A2. This has a responsivity of about 0.9 at 1500 nm, a dark current, i_d, of 10 nA and an excess noise factor of $M^{0.7}$, measured at a multiplication factor, M, equal to 10 [7]. As an example of a good monolithic transimpedance amplifier we take the NE5211D [8] with a frequency response to 180 MHz and a mean square noise current spectral density, S_{th}, of 3.24 (pA)^2/Hz. The total mean square noise current, $\langle i_n^2 \rangle$, referred to the input of the transimpedance amplifier is

$$\langle i_n^2 \rangle = 2eBM^{2.7}(i_d + i_p) + S_{\text{th}}B. \quad (11)$$

Where e is the electronic charge and i_p is the mean (unmultiplied) photocurrent which is the sum of the individual photocurrents generated by the two optical sidebands. To find the minimum value of P_o required, we impose the condition that (owing to the slightly lower CNDR) the ratio $\langle i_{\rm sig}^2 \rangle / \langle i_n^2 \rangle$ must be at least 21 dB to achieve 17.5 dB CNR overall. The optimum value of M is found to be 12 and this gives minimum mean optical powers of -23.5 dBm for the 60 channel system and -31 dBm for 10 channels. Even for large numbers of channels there is some improvement over nonresonant PINFET receivers and for smaller numbers the performance is very close to that of resonant PINFET's. These figures ignore filtering losses which are discussed below.

V. Fabry-Perot Filter Characteristics

The optical filter is must have sufficient bandwidth to pass the modulated sidebands while adequately rejecting other channels. It is desirable that the channels be closely spaced in order to exploit the modulator's bandwidth fully and it is also desirable to minimize filter losses. This involves compromises which are system dependent and we shall address these issues by considering a video distribution system and estimating the performance for various filter parameters.

Narrowband optical filters rely on multiple beam interference and have periodic responses. This imposes the restriction that the filters free spectral range (FSR) must be such that when any channel is selected, no other channel is passed. As an example we will consider a modulator with a good but currently attainable bandwidth of 10 GHz. It is usual to restrict the channels to a 1 octave band to avoid third order intermodulation products and so we consider the range from 5 to 10 GHz. In this case, the minimum FSR is 10 GHz (equal to the modulator bandwidth) as shown in Fig. 1. The 5 and 10-GHz subcarrier channels are unused and form guardbands. The simplest filter is a single FPF. Expressed in terms of the FSR and the full width at half maximum (FWHM), the filter response (neglecting insertion losses) is

$$\frac{P_T}{P_I} = \left[I + \left(\frac{2\mathrm{FSR}}{\pi \mathrm{FWHM}}\right)^2 \sin^2\left(\frac{\pi \Delta f}{\mathrm{FSR}}\right)\right]^{-1} \quad (12)$$

where P_I is incident optical power, P_T is transmitted power, and Δf is frequency offset from the passband center. For $\Delta f \ll$ FSR this is approximated with good accuracy by

$$\frac{P_T}{P_I} = \left[1 + \left(\frac{2\Delta f}{\mathrm{FWHM}}\right)^2\right]^{-1}. \quad (13)$$

If the modulated subcarrier frequency is offset from the unmodulated subcarrier frequency by, say, 20 MHz then the received (30 MHz bandwidth) signal will be 20 MHz $\pm$ 15 MHz. The filter needs to pass modulation frequencies from 5 to 35 MHz without excessive attenuation and with no more FM-AM conversion and distortion than can be compensated for electronically. A reasonable value is around 3 dB signal attenuation (1.5 dB optical attenuation) at 35 MHz and this requires a FWHM of 70 MHz (since the received signal is proportional to the product of the sideband amplitudes). Thus the finesse (= FSR/FWHM) of the FPF is 143, which is readily achievable in a fiber pigtailed device [9], [10]. In general, insertion loss increases with finesse. A finesse of 1000 for an insertion loss of 4 dB has been reported [10], while an earlier device [9] with poorer mirrors had 1.5 dB loss and a finesse of 60; we will assume a value of 1.5 dB.

It only remains to determine the level of crosstalk from other channels which can be tolerated. This contributes to the total noise level and must be included in calculating the overall CNR. Some increase in the minimum received optical power is needed to maintain a CNR of 17.5 dB. A greater sensitivity reduction allows more crosstalk to be tolerated and so closer channel spacing. An exact solution requires the crosstalk level to be included in calculating the optimum modulation index but the effect of this correction is negligible. Instead we simply consider the necessary increase in minimum optical power required. Allowing an increase of 3 dB raises the received signal by 6 dB and allows a 22-dB (electrical) carrier to crosstalk ratio. Most crosstalk arises from the two channels closest to a FPF response peak with progressively smaller contributions from more distant channel pairs. Since crosstalk is dominated by channels in the region where the FPF response falls off at close to 6 dB/octave (and received signal power by 12 dB/octave), we approximate the total crosstalk, C, as

$$C = S - 10 \cdot \log\left(2 \sum_{p=1}^{N} \frac{1}{p^4}\right) \quad (14)$$

where S is the supression of the closest channels and N is the number of channel pairs considered. For many channels, this implies that S must be about 25.4 dB (12.7 dB optical suppression) and substituting this value in (13) indicates a minimum channel separation in the region of 146 MHz. The total sensitivity penalty incurred including the filter insertion loss and an equalisation margin is 6 dB (optical).

To improve upon this channel spacing, multipole filtering is needed. This can be achieved by passing the optical signal through two FPF's, or through the same filter twice [11]. It is necessary to prevent interaction between the filters. This may be achieved by attenuation, by separation exceeding the coherence length of the source, or by optical isolation [12], [13], although attenuation is undesirable and the second method is incompatible with high coherence sources. A fourth possibility is a multiple mirror FPF [13], [14]. Three mirrors form two resonant cavities with no cavity between them. Regardless of the particular technique used, the best adjacent channel rejection is realized if the two filters have similar responses so that they contribute equally to the roll-off. Staggering the individual responses allows a flatter passband and steeper roll-off close to the passband [13], [14]. Taking the simplest case of two identical FPF's with their center frequencies aligned, each needs a finesse of 92 and a FWHM of 109 MHz to obtain 70 MHz overall, and the composite filter roll-off is 12 dB per octave. This allows about 98 MHz channel spacing. The disadvantages are greater insertion loss and complex tuning.

VI. Summary

In this paper we have addressed for the first time the fundamental issues of sensitivity and channel spacing for optically prefiltered SCM systems and compared this with conventional SCM techniques. To achieve the best performance with optical prefiltering requires different modulation methods and external modulator biasing than in conventional systems. The proposed sub-subcarrier modulation scheme may be implemented without major changes to the transmitter and a single FPF can select between many FM video channels. Even including filter losses, sensitivities equal to nonresonant PINFET receivers are possible using existing technology. There are two main drawbacks; the separation required between channels is greater than that needed with electrical filtering and the sensitivity falls relative to conventional receivers for large numbers of channels. This is fundamental to the optically prefiltered approach. Optical prefiltering is, therefore, most suitable for applications where small numbers of wideband channels are required. In such cases considerable receiver bandwidth reduction is possible. A further advantage of prefiltering is insensitivity to fiber dispersion. As discussed in Section II, SCM modulation produces components on either side of the optical carrier at multiples of the subcarrier frequencies which mix and cancel at the receiver. Even slight dispersion causes intermodulation by introducing a phase shift between components so that cancellation is incomplete [15]. The bandwidth over which dispersion acts may be many GHz. In optically prefiltered systems, each channel occupies only the baseband bandwidth and this intermodulation source is removed.

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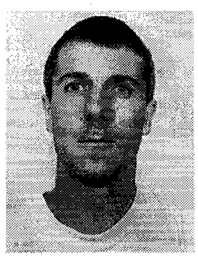

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