Optical Transmission of OFDM Ultra-wideband Signals beyond 40 Gb/s

(Invited Paper)

Y. Ben-Ezra\textsuperscript{2}, M. Ran\textsuperscript{2}, B.I. Lembrikov* \textsuperscript{2}, U. Mahlab \textsuperscript{1,2}, M. Haridim\textsuperscript{2}, A. Leibovich\textsuperscript{2}

\textsuperscript{1}Optical Networks Division, ECI Telecom Ltd., IL-49517, 30 Hasivim Street, Petah-Tikva, Israel,
\textsuperscript{2} Department of Electrical Engineering, Holon Institute of Technology,
P.O. Box 305, 58102, 52 Golomb str., Holon, Israel,

Abstract—We for the first time propose the highly efficient method of RF and optical signal mixing based on two different architectures: the parallel-RF/serial-optics architecture characterized by all-optical mixing for sub-carrier multiplexing, and the parallel-RF/parallel-optics architecture based on the array of 12x10 GHz components with directly modulated VCSELs and 12 multimode optical fibers. The main advantages of both architectures are simplicity and low-cost implementation. We have carried out numerical simulations of ultra-wideband signals propagation in the proposed systems and proved the high efficiency and feasibility of the proposed method.

Index Terms—100 GbE, OFDM, ultra-wideband signals.

I. INTRODUCTION

In optical subcarrier multiplexed (SCM) systems, several channels are multiplexed around an optical carrier using frequency division multiplying (FDM) technique [1], [2]. Each channel can be digitally intensity modulated into a bit-stream or coded using a multilevel modulation scheme at the radio frequency (RF) domain, enabling an increase of the channel information capacity. The frequency multiplexed signal is then converted to an analog optical signal using a very wide optical modulator. In order to correctly transpose the frequency multiplexed signals, a high degree of linearity and a very wide RF bandwidth are required. One of the main drawbacks of SCM technique is the RF fading effect coming from interaction between the RF subcarrier and the chromatic dispersion of the fiber. During the propagation of the double side band signal through a dispersive optical fiber, the upper and the lower side bands will undergo different phase shifts due to the different group velocities. The squared photodetection applied to this double side band signal will exhibits the RF fading effect at determined fiber distances. Dispersion induced RF fading effect is significant in optical links without dispersion compensation. In order to improve the spectral efficiency and to reduce the chromatic dispersion penalties, two subcarrier modulation schemes can be used: single-side band modulation (SSB) and tandem single side band modulation (TSSB) [3]. These two techniques can be realized by using a dual drive Mach-Zehnder modulator (MZM) [3]. However, MZM induces nonlinear signal distortions.

In this paper, we investigated theoretically and experimentally the three aspects of Orthogonal Frequency Division Multiplexing (OFDM) technology:

1) the parallel and serial architecture;
2) the optical link;
3) a novel component for the UWB MB OFDM signal detection.

OFDM technology is a key building block for mitigation of the non-linear signal distortion. Parallel and serial architectures are explored in order to construct multi-band (MB) OFDM signals capable of delivering a multi-gigabit analog signal. The basic element of the both proposed architectures is the optical link consisted of a directly modulated vertical cavity surface emitting laser (VCSEL), multimode optical fiber (MMF) and a p-i-n photodiode (PD). Transmission performances of such a system are analyzed theoretically. We have carried out the numerical simulations of UWB MB OFDM signal transmitted over this optical link. The measurements and simulations have been carried out for the time frequency code 5 (TFC5), TFC6 and TFC7 bands of MB OFDM UWB signals determined by the frequency intervals (3168 ÷ 3696) MHz, (3696 ÷ 4224) MHz, and (4224 ÷ 4752) MHz, respectively [2]. We address beyond 40Gb/s data rates by parallel transmission over more than 128 conventional WiMedia/ECMA [2] baseband channels, each having 528 MHz. One of the key advantages of the proposed approach is the ability to provide hybrid fiber-wireless solution, where the wireless segment at the available ultra-wideband (UWB) transmission is fully compliant with UWB regulations. We proposed a novel microwave photonic component based on SiGe technology in order to realize the low-cost and highly integrable systems with the required performance characteristics. We developed a physical model and carried out numerical simulations of an optically controlled microstrip (MS) convertor (OCMC) based on Si and SiGe on Si (SiGe/Si) structures. The numerical estimations based on the proposed model of the thin layer SiGe/Si OCMC structure with an detecting layer thickness of about \( d = (0.5 ÷ 2) \mu m \) clearly show that a bandwidth of at least \( 60 GHz \) can be achieved.

The paper is constructed as follows. In Section 2 we describe the two types of architecture. In Section 3 the theoretical model of the basic optical link and the wireless channel is discussed. In Section 4 the model of novel OCMC component...
based on SiGe/Si structure is analyzed theoretically. The numerical simulation results for MB OFDM UWB signal transmission are presented in Section 5. The experimental results are discussed in Section 6. Conclusions are presented in Section 7. The details of mathematical derivations are presented in Appendix.

II. TYPES OF ARCHITECTURE

A. Parallel RF/Serial Optic Architecture

A novel concept for a scalable radio-over-fiber (ROF) system enable to bring up to $61.44 \text{Gb/s}$ is shown in Fig. 1. The system is scalable in such a way that it enables the various channels and bands. The development of novel O/E and E/O components and subsystems for the extended band UWB signal transmission over the fiber is necessary. For instance, photo detection up to $64 \text{GHz}$ may be achieved through the lateral illumination and resonant-cavity-enhancement of SiGe heterojunction phototransistor (HPT). Additionally, an ultra-wideband and highly linear E/O modulator is needed for the implementation of the proposed architecture. A single-mode fiber (SMF) can be used in long-haul applications.

B. Beyond 40 Gb/s Parallel RF/Parallel Optic Architecture

In the alternative scheme of the parallel RF/parallel optics architecture shown in Fig. 2 each directly modulated low-cost multimode VCSEL with a 10Gb/s bandwidth transmits its signal over a separate MMF. This architecture based on 12x10 GHz transceiver for digital 100GbE was proposed in [4]. We enhanced this architecture for ROF applications. In contrast to the parallel RF/serial optics architecture with SMF suitable for long-haul applications, this version based on MMF is appropriate for short-range applications. The parallel RF/parallel optics architecture is expected to operate at the wavelength of $850 \text{nm}$ over MMF of the length of about several hundred of meters. The input lanes are directly connected to 12 Laser Drivers (LDs), which are in turn connected to a 12-element VCSEL array. The output lanes are directly connected to a 12-element p-i-n diode array. After the detection the RF signals are amplified by 12 transimpedance amplifiers (TIAs) as it seen from Fig. 2.

III. THEORETICAL MODEL OF AN OPTICAL LINK

The both proposed architectures are based on the optical link as it was mentioned in Section 1. The proposed link shown in Fig. 3 consists of an optical link and wireless channels. In this section we consider existing theoretical models of the optical link containing the directly modulated VCSEL as a transmitter, MMF, p-i-n PD, and a wireless channel. MMF link models have been discussed in a number of works [5]-[8].

Consider the standardized $850\text{nm}$ laser-optimized $50\mu m$ MMF model [5]. Transmitter, MMF and connections are the most important factors determining the $3\text{dB}$ optical bandwidth of the link [5]. The transmitter, VCSEL is a key device in local area networks using MMFs [9]. The VCSEL’s well known advantages are following: low power consumption; high-speed modulation with low driving current; narrow circular beam for direct fiber coupling; low cost and small packaging capability; single longitudinal mode operation with vertical microcavity
The operational characteristics of the directly modulated VCSEL are described by the rate equations for the photon density \( P(t) \), electron density \( N(t) \) and the phase \( \phi(t) \) since the amplitude modulation in semiconductor lasers is accompanied by the phase modulation determined by the linewidth enhancement factor (LEF) \( \alpha_c \) [5], [10]-[13].

\[
\frac{dP}{dt} = \left[ \frac{\Gamma a(N - N_0)}{(1 + \varepsilon P)} - \alpha_t \right] v_g P - \frac{\Gamma N}{\tau_p} + \frac{\beta \Gamma N}{\tau_e} + F_P(t) \tag{1}
\]

\[
\frac{dN}{dt} = \frac{I(t)}{qV} - \frac{v_g a(N - N_0)}{(1 + \varepsilon P)} P - \frac{N}{\tau_e} - B N^2 - C N^3 + F_N(t) \tag{2}
\]

\[
\frac{d\phi}{dt} = \frac{1}{2} \alpha_c \left[ \Gamma v_g a(N - N_0) - \frac{1}{\tau_p} \right] + F_\phi(t) \tag{3}
\]

where \( a \) is the differential gain, \( N_0 \) is a transparency electron concentration, \( \Gamma \) is the confinement factor, \( v_g \) is the group velocity of light, \( V \) is the active region volume, \( \tau_{p,e} \) are the photon and electron lifetimes, respectively, \( \varepsilon \) is the gain compression factor, \( \beta \) is the spontaneous emission fraction coupled into a lasing mode, \( q \) is the electron charge, \( I(t) \) is the VCSEL bias current, \( B \) is the bimolecular recombination factor, \( C \) is the Auger recombination factor, \( \alpha_t \) is the total loss coefficient given by

\[
\alpha_t = \alpha_{\text{loss}} + \frac{1}{L} \ln R \tag{4}
\]

\( \alpha_{\text{loss}} \) is the VCSEL absorption coefficient, \( L \) is the VCSEL active region length, and \( R \) is the reflectivity of the mirrors. The terms \( F_P(t), F_N(t), F_\phi(t) \) are the Langevin forces assumed to be the Gaussian random processes with zero expectancy and the following correlation function

\[
\langle F_i(t) F_j(t') \rangle = 2D_{ij} \delta(t - t') \tag{5}
\]

where \( i, j = P, N, \phi \), the angle brackets denote the ensemble average, and \( D_{ij} \) is the diffusion coefficient in the Markovian approximation. The main contribution to the laser noise is caused by the diffusion coefficients \( D_{PP} \) and \( D_{\phi\phi} \) given by

\[
D_{PP} = \frac{\Gamma v_g a_0(N - N_0)}{V} P; \quad D_{\phi\phi} = \frac{\Gamma v_g a_0(N - N_0)}{VP} \tag{6}
\]

Single mode rate equations (1)-(3) have been found to be a very good approximation to the large signal behavior for MMF [5]. For the analogous applications, the relation between the SNR and the relative intensity noise (RIN) is given by

\[
\text{SNR} = \frac{m^2}{2 RIN} \tag{7}
\]

where \( m = \Delta I / (I - I_{th}) \) is the electrical modulation depth, \( I, I_{th} \) are the input and the threshold currents, respectively. RIN is given by

\[
RIN = \left( \frac{\langle \delta P_{\text{opt}}^2(t) \rangle}{P_{\text{opt}}(t)} \right)^2 \tag{8}
\]

where \( \langle \delta P_{\text{opt}}^2(t) \rangle \) is the mean square optical power fluctuation and \( \langle P_{\text{opt}}(t) \rangle \) is the average optical power.

MMF links performance is affected by degradation due to finite rise and fall times at the transmitter and the receiver, the intermodal and intramodal dispersion, and noises specific to MMF links or multimode lasers [5]. The proposed model mainly concentrates on the signal degradation due to the intermodal dispersion because the largest part of the link power budget consumption is caused by pulse spreading caused by the intermodal dispersion [5]. At the operating wavelength \( \lambda = 850nm \), the 50\( \mu \)m, 1\% \( \Delta \) MMF initially supports 19 mode groups, each of which can have its own group velocity \( v_g \) [5]. In actual MMFs there exists the coupling between the modes due to the fiber imperfections. However, only the coupling of modes within a mode group is significant over the short length scales of hundreds of meters, while the modal dispersion between mode groups is neglected, and the coupling between them is absent [5]. The attenuation of the coupling modes within each group \( \mu \) is described by the attenuation rate \( \gamma_\mu \), and the amplitude of a pulse launched into group \( \mu \) is proportional to the factor \( \exp[\gamma_\mu z] \) as it propagates through MMF [5]. As a result, the bandwidth and the MMF transfer function strongly depend on the excitation conditions determining how much power will be coupled into each mode group, and the signal at the receiver output is determined by the launch conditions, MMF properties, and the link configuration [5].

The transverse modes of a VCSEL are assumed to be the Gaussian beam modes \( u_{pl}(r, \varphi, z, w_0, k) \) centered at the origin \( r = 0 \) of MMF and parallel to the \( z \) axis. They are given by [5]

\[
u_{pl}(r, \varphi, z, w_0, k) = \frac{w_0}{w} \left( \frac{\sqrt{2}}{w} \right)^l I_p \left( \frac{2}{w^2} r^2 \right) \times \exp \left[ -i(kz - \Phi_{pl} + l\varphi) - r^2 \left( \frac{1}{w^2} + \frac{ik}{2R} \right) \right] \tag{9}
\]

where \( p \geq 0, l \geq 0 \) are the radial and angular mode numbers, \( w_0 \) is the spot size at the waist, \( k = 2\pi/\lambda \) is the free space wavenumber, \( I_p \) are the generalized Laguerre polynomials, and

\[
\Phi_{pl}(z, w_0, k) = (2p + l + 1) \arctan \left( \frac{2z}{kw_0^2} \right) \tag{10}
\]

\[
w(z, w_0, k) = w_0 \left[ 1 + \left( \frac{2z}{kw_0^2} \right)^2 \right] \tag{11}
\]

\[
R(z, w_0, k) = z \left[ 1 + \left( \frac{kw_0^2}{2z} \right)^2 \right] \tag{12}
\]

For the few-moded VCSEL the Gaussian beam model is a reasonable approximation [5]. A VCSEL \( u_{pl} \) mode at the air-fiber interface is transformed into a different Gaussian beam mode which then excites the various modes \( \psi_{lmv}(r, \theta) \) of MMF corresponding to the transverse components \( \vec{E}_{z,\varphi} \) of the electric field in the fiber. The modes \( \psi_{lmv}(r, \theta) \) are given by [5]

\[
\psi_{lmv}(r, \theta) = f_{lm}(r) \nu(l \theta) p \tag{13}
\]

where \( l \geq 0 \) and \( m > 0 \) are the eigen-values of the radial and angular parts of (13), the index \( \nu \) denotes angular dependence.
the optical power absorbed in PD, \( P_{\text{abs}} \), is given by [14]

\[
P_{\text{abs}} = \int_{0}^{\infty} \int_{0}^{\infty} r dr d\theta f_{\text{im}}^{2}(r, \nu^{2}(\theta)) = 1
\]  

(14)

The coupling amplitudes \( a_{pl}^{i,mv} \) of the incident Gaussian beam mode with the fiber mode \( \psi_{\text{imvp}}(r, \theta) \) are given by [5]

\[
a_{pl}^{i,mv} = \int_{A} d^{2}x \psi_{\text{imvp}}(x) u_{pl'}(x'; x, w_{0}', k')
\]  

(15)

where the integration is carried out over the area \( A \) of the PD end face. Assuming that impulses from the transmitter induce electric fields at the input end face of MMF have the form \( \sum_{\mu} c_{\mu} r \delta(x', x; w_{0}', k') \delta(t) \) we write the impulse response of MMF \( h(z, t) \) [5]

\[
h(z, t) = \sum_{\mu} w_{\mu} \exp(-\gamma_{\mu} z) \delta(t - t_{\mu} z)
\]  

(16)

where \( c_{\mu} \) are the complex amplitudes, \( w_{\mu} = \sum_{mv} w_{mv} \) is the mode power distribution (MPD), and

\[
w_{mv} = \sum_{pl'} c_{pl'} a_{pl}^{i,mv} \left| a_{pl}^{i,mv} \right|^{2}
\]  

(17)

Consider now a typical p-i-n PD. Its quantum efficiency \( \eta \) is given by [13]

\[
\eta = \frac{P_{\text{abs}}}{P_{\text{opt}}} = \frac{\zeta (1 - r) (1 - \exp(-\alpha_{PD} d))}{1 - r}
\]  

(18)

where \( \zeta \) is p-i-n PD internal quantum efficiency close to unity, \( P_{\text{opt}} \) is the incident optical power at the input of PD, \( P_{\text{abs}} \) is the optical power absorbed in PD, \( r \) is the reflection coefficient of the PD surface, \( \alpha_{PD} \) is the PD material absorption coefficient, and \( d \) is the thickness of the PD absorption intrinsic layer. The p-i-n PD bandwidth \( \Delta f \) is determined by the carrier transit time and time constant of the p-i-n PD equivalent circuit. It is given by [14]

\[
\Delta f = \left[ \frac{2\pi d}{3.5\tau_{d}} \right]^{2} + \left[ \frac{2\pi \varepsilon_{r} \varepsilon_{0} S (R_{s} + R_{l})}{d} \right]^{2} \right]^{-1/2}
\]  

(19)

where \( \tau_{d} \) is the average charge carrier drift velocity in the PD absorption intrinsic layer, \( \varepsilon_{0} \) is the free space permittivity, \( \varepsilon_{r} \) is the PD permittivity, \( S \) is the PD photosensitive area, \( R_{s}, R_{l} \) are the series and load resistances in the PD equivalent circuit, respectively.

UWB wireless channel description is based on the modified Saleh - Valenzuela (SV) model [15]. In this model the analytical representation of a discrete multipath impulse \( h_{i}(t) \) can be presented as follows [15]

\[
h_{i}(t) = X_{i} \sum_{l=0}^{L_{i} - 1} \sum_{k=0}^{K_{i} - 1} \alpha_{kl}^{i} \delta(t - T_{l}^{i} - \tau_{kl}^{i})
\]  

(20)

where \( \alpha_{kl}^{i} \) are the multipath gain coefficients, \( T_{l}^{i} \) are the delays of the \( l \)th cluster, \( \tau_{kl}^{i} \) is the delay for the \( k \)th multipath component relative to the \( l \)th cluster arrival time. Shadowing effect obeys the log-normal distribution and is represented by \( X_{i} \) where \( i \) refers to the \( i \)th realization. The channel coefficients \( \alpha_{kl}^{i} \) are given by [15]

\[
\alpha_{kl}^{i} = p_{kl} \xi_{l} \beta_{kl}
\]  

(21)

where \( \xi_{l} \) is the fading associated with the \( l \)th cluster, and \( \beta_{kl} \) is the fading associated with the \( k \)th ray of the \( l \)th cluster. The cluster arrival time distribution \( p(T_{l} | T_{l-1}) \) and the ray arrival time distribution \( p(\tau_{kl} | \tau_{l-1}) \) are given by, respectively [15]

\[
p(T_{l} | T_{l-1}) = \Lambda \exp[-\Lambda (T_{l} - T_{l-1})], l > 0
\]  

(22)

and

\[
p(\tau_{kl} | \tau_{l-1}) = \lambda \exp[-\lambda (\tau_{kl} - \tau_{l-1})], k > 0
\]  

(23)

where \( \Lambda \) is the cluster arrival rate, and \( \lambda \) is rate arrival rate. The modified SV model derives as an output mean and root mean square (RMS) excess delays, number of multipath components, and power decay profile.

The numerical simulations of the optical link and wireless channel using the models discussed above had been carried out in the framework of the advanced design system (ADS) package, version 2006, update 2, product of Agilent. The package contains the UWB toolbox based on "Multiband OFDM Physical layer Specification", (WiMedia Alliance document, Release 1.1, July 14, 2005).

IV. NOVEL OCMC COMPONENT

In the case of the parallel RF/serial optic architecture the detection of multiplexed MB OFDM UWB modulated optical signal is required. The output UWB RF signal should be the minimum distorted replica of the original multiplexed MB OFDM UWB envelope of the optical signal at the input to the optical fiber. The optical handling of microwave (MW) devices such as directional couplers, phase shifters, attenuators, ultra-fast MW switches, etc. has been thoroughly investigated both theoretically and experimentally and used successfully in many applications in the framework of Microwave Photonics approach [16]-[26]. The advantages of this approach are following: low cost solution, low power consumption, high responsivity, flat spectral response over the desired band, low noise characteristics, possibility of creation of compact components which can be easily integrated with other electronic and photonics systems [17].

However, the conditions of the detection process are essentially different from the steady state optical control of the MS load. The input signal of the system in our case is the multiplexed MB OFDM UWB modulated optical radiation fed from an optical fiber. Typically, the optical carrier power \( P_{\text{opt}} \) is comparatively low: \( P_{\text{opt}} \sim 1\mu W \). For a multimode optical fiber with an optical beam radius \( r_{b} \sim 10 \mu m \), it yields a comparatively low intensity \( I = P_{\text{opt}} / (\pi r_{b}^{2}) \sim 3 W/mm^{2} \). Instead of it, the detected UWB RF signal voltage over the optically controlled load should serve itself as a source of the MW radiation. We propose a novel OCMC device for this purpose. In the proposed method we used OCMC consisting of an open ended microstrip (MS) line with a semiconducting substrate, as sketched in Fig. 4. The optical beam, modulated
by UWB RF signal, is illuminated on the substrate near the open end of the MS line.

The down conversion from the optical domain to the MW domain can be modeled by an optically controlled load connected at the open end of the MS line. The variations of the photocurrent at the optically controlled load of the MS produce an electromagnetic (EM) waves that propagate along the MS line towards the output port of OCMC from which they are probed by a coaxial line of the same characteristic impedance, $Z_0$. The efficiency of the optical-microwave frequency down conversion depends on the ability to collect the photocarriers at the bottom contact. In the case of thin Ge-on-Si, SiGe/Si, or Si layers of thickness ranges from 350 – 500μm which is quite large compared to the diffusion length of the photocarriers $L_{n,p} = \sqrt{D_{n,p} \tau}$ ~ (10 ÷ 30) μm. In the case of surface absorption characterized by large values of absorption coefficient $\alpha$ and consequently a very small absorption length $\sim \alpha^{-1}$ the effective depth the photocarriers can reach is determined by the diffusion and drift properties of the photocarriers. The feasibility of the proposed OCMC device was experimentally verified by an open-ended MS line with $Z_0 = 50\Omega$ implemented on a high resistivity $\rho > 3000\Omega cm$ slightly p-type Si substrate are shown in Fig. 4. The optical source was a tunable laser diode with wavelengths from $\lambda = 680$ up to $\lambda = 980nm$. The results for the OCMC response function at the different levels of the optical power are shown in Fig. 5. These results do not satisfy the requirements of the UWB RF signal detection.

An alternative approach has been proposed recently in a number of works [27]-[30]. It has been demonstrated experimentally that thin Ge-on-Si, SiGe/Si, or Si layers of a thickness about one up to several micrometers can operate successfully as UWB RF signal detectors providing a bandwidth of about (10 ÷ 20) GHz [27]-[30]. A resonant cavity-enhanced Si photodetector permits to overcome the comparatively low absorption in Si by using the substrate with a distributed Bragg reflector (DBR) that provides 90% reflection of an optical power back into the detector layer [27]. Silicon photodetectors monolithically integrated with preamplifier circuits have achieved error-free detection at up to 5Gbps at an optical wavelength $\lambda = 850nm$ [30]. For operation at longer wavelengths Ge-on-Si photodiodes with the bandwidth up to 21 GHz at $\lambda = 1.31\mu m$ are attractive for monolithic optical receivers [30]. A theoretical model of such thin film devices has not yet been developed to our best knowledge.

Consider an infinite in the $x,y$ directions layer of a thickness $d$ in the $z$ direction placed on a semi-infinite in the $z$ direction substrate. The geometry of the problem is presented in Fig. 6. The electric and magnetic fields of the incident and reflected waves $E_{1x}, H_{1y}$ in the free space $z < 0$, $E_{2x}, H_{2y}$ in the layer $0 \leq z \leq d$, and $E_{3x}, H_{3y}$ in the substrate $z > d$ are given by [31]

$$z < 0 \rightarrow E_{1x} = [E_{1x}^+ \exp(-i k_1 z) + E_{1x}^- \exp(i k_1 z)] \times \exp(i \omega_{opt} t)$$

$$H_{1y} = \frac{1}{Z_1} [E_{1x}^+ \exp(-i k_1 z) - E_{1x}^- \exp(i k_1 z)] \exp(i \omega_{opt} t)$$

Fig. 4. Schematic view of an optically controlled microstrip convertor (OCMC)

Fig. 5. The dependence of the normalized response function of Si based OCMC on a bandwidth for different values of an optical power. The thickness of OCMC substrate $d = 520\mu m$

Fig. 6. Illuminated SiGe layer on a Si substrate
0 ≤ z ≤ d → E_{2x} = \left[ E_2^+ \exp(-\gamma_2 z) + E_2^- \exp(\gamma_2 z) \right] 
× \exp(i\omega_{opt} t) \tag{26}

\frac{1}{Z_2} \left[ E_2^+ \exp(-\gamma_2 z) - E_2^- \exp(\gamma_2 z) \right] \exp(i\omega_{opt} t) \tag{27}

z > d → E_{3x} = \frac{1}{Z_3} \exp(-i\gamma_3 Z) \exp(i\omega_{opt} t); \tag{28}

\frac{1}{Z_3} E_3^+ \exp(-i\gamma_3 Z) \tag{29}

Here, the wave impedances of the media have the form

\begin{align*}
Z_1 &= \sqrt{\frac{\mu_0}{\varepsilon_0}} = 377\Omega; \ Z_2 = |Z_2| \exp i\beta z; \ Z_3 = \sqrt{\frac{\mu_0}{\varepsilon_0}}
\end{align*} \tag{30}

\mu_0, \varepsilon_0 \text{ are the free space permeability and permittivity, respectively, the absorption layer wave impedance } Z_2 \text{ is assumed to be complex, } \varepsilon_3 \text{ is the permittivity of the substrate, } \omega_{opt} \text{ is the optical frequency,}

\begin{align*}
k_1 &= \frac{\omega}{c}, \gamma_2 = \frac{\alpha}{2} + i\beta, k_3 &= \frac{\omega}{c}\sqrt{\varepsilon_3}
\end{align*} \tag{31}

\beta \text{ is the propagation constant, and } c \text{ is the speed of light in vacuum. The solution of the boundary problem yields the expression for the optical intensity } I_{opt}^z (z) \text{ in the thin film layer with absorption.}

\begin{align*}
I_{opt}^z (z) &= I_0 \left[ 2Z_3 \cos \theta \right. 
\left. \left( 1 + \frac{Z_2^2}{Z_3^2} \right) \right] 
+ \sinh (\alpha (z - d)) \left( 1 + \frac{Z_3^2}{Z_2^2} \right) \tag{32}
\end{align*}

where

\begin{align*}
I_0 &= \frac{2Z_1 P_{opt}}{A_{eff} |D|^2 Z_2} 
\frac{1}{2} \left( \frac{Z_1}{Z_2} \right)^2 A_{eff} = \pi r_b^2
\end{align*} \tag{33}

\begin{align*}
P_{opt} &= \frac{|E_1|^2 A_{eff}}{2Z_1}; \ A_{eff} = \pi r_b^2
\end{align*} \tag{34}

\begin{align*}
|D|^2 &= \left| \sinh (\gamma_2 d) \left( 1 + \frac{Z_1 Z_3}{Z_2^2} \right) + \frac{(Z_1 + Z_3)}{Z_2} \cos (\gamma_2 d) \right|^2 \tag{35}
\end{align*}

\( P_{opt} \) is the optical power of the incident wave in the free space \( z < 0 \), and \( r_b \) is the light beam radius. The explicit expression of \(|D|^2\) in general case is hardly observable, and we do not present it here. It can be substantially simplified under the realistic quasi-resonance assumption for \( \lambda_{opt} \sim 1\mu m \) and \( d \sim (0.5 \div 2) \mu m \)

\begin{align*}
\sin \beta d = 0, \beta d = \pi m, m = 1, 2, ..., \tag{36}
\end{align*}

Then, a simplified expression of \(|D|^2\) takes the form

\begin{align*}
|D|^2 &= \frac{(Z_1 + Z_3)^2}{Z_2^2} \cos^2 \left( \frac{\alpha d}{2} \right) + \sin^2 \left( \frac{\alpha d}{2} \right) 
+ \sinh^2 \left( \frac{\alpha d}{2} \right) \left[ 1 + 2 \frac{Z_1 Z_3 \cos 2\theta}{Z_2^2} + \frac{(Z_1 Z_3^2)}{Z_2^2} \right] 
\tag{37}
\end{align*}

The detailed expression of \((32)\) is presented in Appendix.

Evaluate now the concentration the photocarriers in the framework of the drift-diffusion model \([32] - [34]\). The continuity equations for the photoinduced electron and hole concentration \( n(z,t) \) and \( p(z,t) \) have the form, respectively

\begin{align*}
\frac{\partial n}{\partial t} &= n \mu_e E \frac{\partial n}{\partial z} + D_n \frac{\partial^2 n}{\partial z^2} + g_n(z,t) - \frac{n - n_0}{\tau_n} \tag{38}
\frac{\partial p}{\partial t} &= -p \mu_h E \frac{\partial p}{\partial z} + D_p \frac{\partial^2 p}{\partial z^2} + g_p(z,t) - \frac{p - p_0}{\tau_p} \tag{39}
\end{align*}

\( g_n(z,t) = g_p(z,t) = g(z,t) = \eta \frac{\partial I(z,t)}{\partial z} \tag{40} \]

Then the generation rates of electrons and holes \( g_{n,p}(z,t) = g(z,t) \) can be written as follows

\begin{align*}
g(z,t) &= g_0(z) + g_1(z,t); \tag{42}
g_0(z) &= \frac{\eta}{h\nu} \frac{\partial I_{opt}^z(z)}{\partial z} \tag{43}
g_1(z,t) &= \frac{\eta}{h\nu} \frac{\partial I_{opt}^z(t)}{\partial z} f(t) = g_0(z) f(t) \tag{44}
\end{align*}

where \( \tau_{n,p}, D_{n,p}, \mu_{n,p} \) are the lifetime, diffusion coefficients, and mobilities of electrons and holes, respectively, \( n_0, p_0 \) are the equilibrium electron and hole concentrations, \( \eta \) is a quantum efficiency, \( E \) is the electrostatic field, and \( f(t) \) is the UWB RF envelope of the optical carrier \((32)\). The coordinate averaged Fourier transform \( \overline{N}_1(\omega) \) of the photocarrier concentration time-dependent part \( n_1(z,t) \) is actually the OCMC response function. It has the form

\begin{align*}
\overline{N}_1(\omega) &= \frac{1}{d} \int_0^d N_1(z,\omega) dz = \frac{F(\omega) \tau (1 - i\omega\tau)}{d(\alpha^2 L_{aeq}^2 - 1)} \left[ 1 + (\omega \tau)^2 \right] 
\times \frac{1}{\alpha} \left[ I_{01}(\cos(\alpha d) - 1) + I_{02}(\sinh(\alpha d)) \right] 
\tag{45}
\end{align*}

The detailed derivation of \( \overline{N}_1(\omega) \) is presented in Appendix II. The results of the numerical evaluations of the response function \( \overline{N}_1(\omega) \) for the typical values of material parameters of SiGe/Si and Si are presented in Figs.7, 8. The power absorption coefficient of \( \alpha \) \( Si_{1-x}Ge_{x} \) compounds \( \alpha \sim 10^5 cm^{-1} \) in the interval of \( \lambda_{opt} \sim 850 nm \). The electron mobility reaches
For smaller concentrations of Ge the charge carrier mobilities are closer to the ones of a pure Ge, while in the opposite case they tend to the values of a pure Si charge carrier mobilities.

The numerical estimations based on the proposed analytical model of the thin layer SiGe/Si OCMC structure with an detecting layer thickness of about \( d = (0.5 \div 2) \mu m \) clearly show that a bandwidth of at least 60GHz can be achieved as it is seen from Fig. 7. The resonant conditions (36) are essential for the layer thickness because the reflection from the SiGe/Si interface in such a case reaches its maximum value. The proposed structure is simpler as compared to resonant-cavity-enhanced (RCE) photodetectors with DBR layers in the substrate. Generally, the SiGe/Si structures are promising candidates for the high-speed optoelectronics receivers due to the high operation rate, comparatively optical high absorption coefficient, the possibility of operation in the near IR spectrum from 850nm to 1550nm, low noise and compatibility with Si based electronic components.

V. Simulation Results

The numerical simulations have been carried out for the parallel RF/parallel optics architecture. We investigated the mixing of 10 RF channels each one with the 0.5GHz bandwidth. The resulting signal was applied to the multimode 10GHz VCSEL, the modulated optical signal was transmitted through the 50m MMF and at the output detected by the p-i-n PD. The simulation results are shown in Fig. 9. The mixed RF spectrum at the VCSEL input, the modulated optical signal at the VCSEL output, and the detected RF spectrum are shown in the upper box, the middle box, and the lower box of Fig. 9, respectively. The internal structure of one of the RF channels located at 3.5GHz central frequency at the corresponding transmission stages is shown in Fig. 10. This channel includes 128 subcarriers and is transmitting 496Mb/s over 0.5GHz bandwidth. In order to study the dispersion influence on the quality of the transmitted MB OFDM signals we have carried out the simulation for the different MMF lengths. The short MMF with a length of 50m has an almost flat frequency response up to the frequency of 10GHz. The
Fig. 11. The calculated magnitude (the upper box) and the phase (the lower box) of the 650 m MMF transfer function

strongly inhomogeneous behavior in such a case in the vicinity of \(10\, \text{GHz}\) is caused by the VCSEL bandwidth limitations. The p-i-n PD used in these measurements has the bandwidth of about \(25\, \text{GHz}\). The bandpass filter behavior of the MMF caused by the multimode dispersion is strongly manifested for longer MMFs. The magnitude and the phase of the 650 m MMF are shown in Fig. 11. In the TFC7 frequency range the transfer function magnitude curve is flat. On the contrary, in the TFC6 frequency range the transfer function magnitude has a notch. The transmission of the multiplexed MB OFDM signals is limited by the MMF length of about 100 m.

VI. THE EXPERIMENTAL RESULTS

A. MB OFDM Signal Transmission

MB OFDM UWB signal was directly applied to the VCSEL and after the propagation through the MMF was detected by the p-i-n PD. The objective of the measurements was to study the performance of the proposed link by means of the packet error rate (PER). The measurements have been carried out for the TFC5, TFC6 and TFC7 band of MB OFDM UWB signals. Fig. 12 presents the PER dependence versus MMF length for the optical link only. The PER dependence versus the MMF length for the different MB OFDM UWB signals shows a peculiar behavior. The PER of the TFC7 band located at higher carrier frequency (4.488 GHz) stays flat and has values of an order of magnitude of \(10^{-6}\) for MMF length up to 1 km. However, the PER in the case of TFC5 band located at the 3.5 GHz carrier frequency and the TFC6 band located at the 4.0 GHz carrier frequency increases dramatically for the MMF lengths longer than 300 m.

In order to understand this behavior of the PER versus MMF length we have measured the MMF transfer function for different MMF lengths. The short MMF with a length of 10 m has an almost flat frequency response up to the frequency of 10 GHz as it is shown in Fig. 13. The strongly inhomogeneous behavior in such a case in the vicinity of 10 GHz is caused by the VCSEL bandwidth limitations. The p-i-n PD used in these measurements has the bandwidth of about 25 GHz. The bandpass filter behavior of the MMF caused by the multimode dispersion is strongly pronounced for longer MMFs.

According to Fig. 13, the MMF transfer function has strong notches in the frequency range of TFC5 band at the fiber lengths longer than 500 m. These strong notches affect the signal spectrum and lead to the significant increase of the PER. PER versus MMF length for the combined MMF and wireless link is shown in Fig. 14. In these measurements carried out for the different MMF lengths we kept constant the distance between two antennas of the wireless channel. These experimental results are in good accord with the simulations results mentioned above.

B. All-optical Up-conversion of MB OFDM Signals

For the parallel RF/series optics architecture shown in Fig. 1 all-optical up-conversion can be applied instead of the conventional RF up-conversion. The low cost all-optical up-conversion is realized using VCSEL’s nonlinearity. In such a case, VCSEL is biased simultaneously by a local oscillator (LO) and UWB signal. The second and third-order intermodulations are situated in frequency ranges \(f_{LO} \pm f_{UWB}, 2f_{LO} \pm f_{UWB}, f_{LO} \pm 2f_{UWB}\) where \(f_{LO}, f_{UWB}\) are the LO and the UWB central frequencies, respectively. The second-order intermodulation products are dominant in the vicinity of the VCSEL threshold bias current. However, due to the large bandwidth of UWB OFDM signal the frequency \(f_{UWB} + f_{LO}\)
Fig. 14. PER dependence on the MMF length for the combined optical and wireless link

Fig. 15. The constellation and spectrum of the transmitted MB OFDM signal with a central frequency 6.6GHz may fall into the up-converted UWB signal spectrum and for this reason it cannot be singled out. This bandwidth overlapping limits the possibility of the up-conversion for a wide range of the UWB OFDM signal frequencies.

On the contrary, in the case of the third order nonlinearity the distortion terms of the type $2f_{UWB} + f_{LO}$ and $2f_{LO} + f_{UWB}$ fall outside the UWB OFDM signal bandwidth. For this reason, it is possible to provide a much better performance of the up-conversion of a wide range of UWB signal frequencies $f_{UWB}$ by means of the third-order intermodulation. We used the UWB signal power of $P_{UWB} = -14dBm$, the LO power of $-5dBm$, and the bias current of 3mA.

The measured error-free constellation diagrams and undistorted spectra of the both up-converted UWB signals for TFC5, TFC6, and TFC7 bands with the central frequencies $f_{UWB} = 6.6GHz$ and $f_{UWB} = 7.128GHz$, respectively, are shown in Figs. 15, 16. These results prove the feasibility of the proposed low cost all-optical up-conversion.

VII. CONCLUSIONS

In conclusion, we proposed two possible architectures for the high spectral efficiency optical transmission of OFDM UWB signals beyond 40Gb/s: the parallel RF/serial optics architecture and parallel RF/parallel optics architecture. We have carried out the numerical simulations for the parallel RF/parallel optics architecture and predicted its highly quality performance. We investigated theoretically and experimentally the optical link consisted of the directly modulated VCSEL, MMF, p-i-n PD and a wireless channel. We presented the detailed theoretical analysis and numerical results for a novel OCMC detecting device based on the SiGe/Si structure. We demonstrated experimentally the highly efficient and low cost all-optical up-conversion of UWB signals.

VIII. ACKNOWLEDGEMENTS

This work was supported in part by
1) The European Project UROOF-Photonic Components for UWB over Optical Fiber (IST-5-033615);
2) The Joint HIT-ECI DIAMOND Project supported by the Office of the Chief Scientist of the Israel Ministry of Industry.
APPENDIX

EVALUATION OF THE OPTICAL INTENSITY IN A THIN LAYER WITH ABSORPTION

The boundary conditions for the electric and magnetic fields at the layer surfaces \( z = 0 \) and \( z = d \) yield [31]:

\[
z = 0 \rightarrow E^+_1 + E^+_2 = E^+_2 + E^-_2
\]

\[
\frac{1}{Z_1} [E^+_1 - E^-_1] = \frac{1}{Z_2} [E^+_2 - E^-_2]
\]

\[
z = d \rightarrow E^+_2 \exp(-\gamma_d d) + E^-_2 \exp(\gamma_d d) = E^+_1 \exp(-ik_3d)
\]

\[
\frac{1}{Z_2} [E^+_2 \exp(-\gamma_d d) - E^-_2 \exp(\gamma_d d)] = \frac{1}{Z_3} E^+_3 \exp(-ik_3d)
\]

We assume that the incident wave amplitude \( E^+_1 \) is known. Then, eliminating \( E^+_1 \) and \( E^+_2 \) we obtain

\[
E^+_2 = \frac{E^+_1 \exp(\gamma_d d) \left[ 1 + \frac{Z_2}{Z_3} \right]}{\left[ \sinh(\gamma_d d) \left[ 1 + \frac{Z_3 Z_2}{Z_2^2} \right] + \frac{(Z_3 + Z_2)}{Z_2^2} \cosh(\gamma_d d) \right]}
\]

\[
E^- = \frac{E^+_1 \exp(-\gamma_d d) \left[ 1 - \frac{Z_2}{Z_3} \right]}{\left[ \sinh(\gamma_d d) \left[ 1 + \frac{Z_3 Z_2}{Z_2^2} \right] + \frac{(Z_3 + Z_2)}{Z_2^2} \cosh(\gamma_d d) \right]}
\]

Substituting (50) and (51) into (26) and (27) we obtain the expressions for the electric and magnetic field in the layer.

\[
E^+_2(x, z, t) = \frac{E^+_1 \exp(-\gamma_2 (z - d)) \left[ 1 + \frac{Z_2}{Z_3} \right]}{\left[ \sinh(\gamma_2 d) \left[ 1 + \frac{Z_3 Z_2}{Z_2^2} \right] + \frac{(Z_3 + Z_2)}{Z_2^2} \cosh(\gamma_2 d) \right]} \exp i\omega t
\]

\[
E^- = \frac{E^+_1 \exp(\gamma_2 (z - d)) \left[ 1 - \frac{Z_2}{Z_3} \right]}{\left[ \sinh(\gamma_2 d) \left[ 1 + \frac{Z_3 Z_2}{Z_2^2} \right] + \frac{(Z_3 + Z_2)}{Z_2^2} \cosh(\gamma_2 d) \right]} \exp i\omega t \quad \text{(52)}
\]

\[
H^+_2(y, z, t) = \frac{E^+_2(x, z, t)}{Z_2} \quad H^-_2(y, z, t) = -\frac{E^-_2(z, t)}{Z_2} \quad \text{(53)}
\]

The time averaged total optical intensity \( I_{opt}^{tot} \) in the layer consists of the intensities \( \langle P^+ \rangle \) and \( \langle P^- \rangle \) of the incident wave and reflected wave, respectively. It has the form [31]

\[
I_{opt}^{tot}(z) = \langle P^+ \rangle + \langle P^- \rangle \quad \text{(55)}
\]

where

\[
\langle P^+ \rangle = \text{Re} \left\{ \frac{1}{2} E^+_2(x, z, t) \left( H^+_2(y, z, t) \right)^* \right\}
\]

\[
\langle P^- \rangle = \text{Re} \left\{ \frac{1}{2} E^+_2(x, z, t) \left( H^-_2(y, z, t) \right)^* \right\}
\]

\[
= \text{Re} \left\{ \frac{|E^+_2|^2 \exp(-2\Re(\gamma_2)(z - d)) \left[ 1 + \frac{Z_2}{Z_3} \right]}{2Z_2^2 \sinh(\gamma_2 d) \left[ 1 + \frac{Z_3 Z_2}{Z_2^2} \right] + (Z_1 + Z_3) \cosh(\gamma_2 d) \right\}^2 \right\}
\]

\[
\langle P^- \rangle = \text{Re} \left\{ \frac{1}{2} E^-_2(x, z, t) \left( H^-_2(y, z, t) \right)^* \right\}
\]

\[
= \text{Re} \left\{ \frac{|E^+_2|^2 \exp(2\Re(\gamma_2)(z - d)) \left[ 1 - \frac{Z_2}{Z_3} \right]}{2Z_2^2 \sinh(\gamma_2 d) \left[ 1 + \frac{Z_3 Z_2}{Z_2^2} \right] + (Z_1 + Z_3) \cosh(\gamma_2 d) \right\}^2 \right\}
\]

Taking into account that according to (31) \( 2\Re(\gamma_2) = \alpha \) and substituting (56) and (57) into (55) we finally obtain expression (32).

II.

APPENDIX

EVALUATION OF THE PHOTOINDUCED CARRIER CONCENTRATION

Substituting (32) into equation (43) we obtain

\[
g_0(z) = I_{01} \sinh(\alpha(z - d)) - I_{02} \cosh(\alpha(z - d)) \quad \text{(58)}
\]

where

\[
I_{01} = \frac{\eta \alpha}{h \nu} \int_0^{2Z_3} \cos \theta \; Z_2 \; I_0 \left[ 1 + \frac{Z_3^2}{Z_2^2} \right] \quad I_{02} = \frac{\eta \alpha}{h \nu} I_0 \left[ 1 + \frac{Z_3^2}{Z_2^2} \right] \quad \text{(59)}
\]

Typically, in the photoinduced plasma, the electron and hole relaxation time \( \sim 10 \text{ns} \) is much smaller than the carrier lifetime, and electroneutrality condition can be applied [22], [23], [34].

\[
n = p \quad \text{(60)}
\]

At the illuminated surface of the semiconductor the strong injection mode and ambipolar diffusion are realized when \( n \gg n_0, p_0 \) and the ambipolar mobility \( \mu_n \) vanishes [20], [21]

\[
\mu_n = \frac{\mu_n p_0(p - n)}{\mu_p p + \mu_n n} = 0 \quad \text{(61)}
\]

In our case the thin layer is entirely occupied by the strong injection mode region. Under such conditions continuity equations (38), (39) reduce to the ambipolar diffusion equation [34]

\[
\frac{\partial n}{\partial t} = D_a \frac{\partial^2 n}{\partial z^2} - \frac{n}{\tau} + g(z, t) \quad \text{(62)}
\]

where it is assumed that \( \tau_n = \tau_p = \tau \). According to expressions (42)-(44) we separate the steady-state and time dependent parts \( n_{ph}(z, t) \), \( n_1(z, t) \) of the photocarrier concentration \( n \).

\[
n = n_{ph}(z) + n_1(z, t) \quad \text{(63)}
\]

Substituting (63) into equation (62) we obtain two equations for \( n_{ph}(z) \) and \( n_1(z, t) \), respectively

\[
D_a \frac{\partial n_{ph}}{\partial z} - \frac{n_{ph}}{\tau} + g_0(z) = 0 \quad \text{(64)}
\]

\[
\frac{\partial n_1}{\partial t} = D_a \frac{\partial^2 n_1}{\partial z^2} - \frac{n_1}{\tau} + g_0(z) f(t) \quad \text{(65)}
\]
where the ambipolar diffusion coefficient $D_a$ is given by [20]

$$D_a = \frac{2D_nD_p}{(D_a + D_p)}$$  \hspace{1cm} (66)

We use the boundary conditions of the mixed type. We assume a finite surface recombination rate $s_0$ on the top surface $z = 0$ which yields [23]

$$\frac{\partial n}{\partial z} |_{z=0} = \frac{s_0}{D_a} n (z = 0)$$  \hspace{1cm} (67)

On the other hand, a kind of an ohmic contact at the interface of the layer and the substrate [27] $z = d$ prescribes the condition

$$n (z = d) = 0$$  \hspace{1cm} (68)

We are interested in the time-dependent part of the photocarrier concentration $n_1 (z,t)$ which is responsible for the UWB RF signal detection. Hence we should solve equation (65) with the boundary conditions (67) and (68). In general case of the UWB RF signal $f (t)$ we carry out the Fourier transform of equation (65) with respect to time. We obtain

$$D_a \frac{\partial^2 N_1 (z, \omega)}{\partial z^2} - (i \omega + \frac{1}{\tau}) N_1 (z, \omega) + g_0 (z) F (\omega) = 0$$  \hspace{1cm} (69)

where

$$N_1 (z, \omega) = \int_{-\infty}^{\infty} n_1 (z, t) \exp (-i \omega t) \, dt$$  \hspace{1cm} (70)

and

$$F (\omega) = \int_{-\infty}^{\infty} f (t) \exp (-i \omega t) \, dt$$  \hspace{1cm} (71)

The boundary conditions (67) and (68) can be applied to the general solution of (69). The result has the form.

$$N_1 (z, \omega) = \frac{F (\omega) \tau (1 - i \omega \tau)}{(\tau^2 L_{aeq}^2 - 1) [1 + (\omega \tau)^2]}$$

$$\times \left[ -g_0 (z) + \frac{1}{L_{aeq}} \cosh \left( \frac{d}{L_{aeq}} \right) + \frac{s_0}{D_a} \sinh \left( \frac{d}{L_{aeq}} \right) \right]$$

$$\times \left[ \frac{\partial g_0}{\partial z} (0) - \frac{s_0}{D_a} g_0 (0) \right] \sinh \left( \frac{z - d}{L_{aeq}} \right)$$

$$+ g_0 (d) \left[ \frac{1}{L_{aeq}} \cosh \left( \frac{z}{L_{aeq}} \right) + \frac{s_0}{D_a} \sinh \left( \frac{z}{L_{aeq}} \right) \right]$$  \hspace{1cm} (72)

where

$$L_{aeq}^2 = \frac{D_a \tau (1 - i \omega \tau)}{1 + (\omega \tau)^2}$$  \hspace{1cm} (73)

The expression (72) for $N_1 (z, \omega)$ averaged over the layer thickness $d$ can be used as the frequency response of the illuminated layer when $f (t) = \delta (t)$ and consequently $F (\omega) = 1$. Using the explicit expression (58) for $g_0 (z)$ we obtain expression (45).


