Theoretical and Experimental Investigations of Direct-Detected RF-Tone-Assisted Optical OFDM Systems

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***Abstract*—In this paper, we propose and experimentally demonstrate a radio frequency (RF)-tone-assisted optical orthogonal frequency-division multiplexing (OFDM) transmission. By inserting an RF tone at the edge of the signal band and biasing the Mach–Zehnder modulator (MZM) at the null point, the proposed system has a better sensitivity and chromatic dispersion (CD) tolerance compared to the previous intensity-modulated single-sideband OFDM (SSB-OFDM). We show analytically that the majority of the linear channel impairments, such as the transmitter, CD, optical filtering, and receiver, can be compensated for by a simple zero-forcing equalizer. Besides, the optimum value of the important parameter, carrier-to-signal-power ratio (CSPR), is analytically obtained and supported via the experimental results. We also observe that the relatively worse sensitivity of the previous SSB-OFDM can be attributed to the limited CSPR. We experimentally demonstrate a 10-Gb/s, 8 quadrature-amplitude modulation (QAM) RF-tone-assisted OFDM transmission, and show that our system has a ∼5-dB better sensitivity compared to the previous intensity-modulated SSB-OFDM and exhibits a negligible transmission penalty after 260-km uncompensated standard single-mode fiber (SSMF).**

***Index Terms*—Direct detection, optical fiber communication, optical modulation, orthogonal frequency-division multiplexing (OFDM).**

I. Introduction

OPTICAL orthogonal frequency-division multiplexing (OFDM) has recently gained much attention due to its potential of electrical equalization to mitigate various deleterious effects, such as chromatic dispersion (CD) [1]–[3] and polarization mode dispersion (PMD) [4]–[6]. The optical OFDM system can be mainly categorized as 1) coherent and 2) incoherent (direct-detected) systems. In general, coherent OFDM exhibits better sensitivity and better spectral efficiency (SE) than a direct-detected OFDM system. However, the coherent approach requires a local oscillator, polarization stabilizer, and extra phase and frequency offset estimation at the receiver, thus increasing the complexity of both the transmitter and the receiver. On the other hand, the direct-detected OFDM requires only one photodiode at the receiver and thus is very easy to be implemented. In addition, when combined with the polarization division multiplexing (PDM) and the self-diversity receiving, the SE of the direct-detected approach can be further doubled without using any adaptive polarization controller [7].

Recently, a spectrally efficient direct-detected OFDM format, which is referred to as the intensity-modulated single-sideband OFDM (SSB-OFDM), has been reported which uses either a dual-drive Mach–Zehnder modulator (DD-MZM) [8], [9] or an optical in-phase/quadrature-phase (I/Q) modulator [10], [11]. However, that technique in [8] and [9] has a design tradeoff between better sensitivity and robustness to fiber CD. Thus, it would be quite advantageous to have an alternative OFDM format that exhibits a good receiving sensitivity while still keeps a good CD tolerance.

In this paper, we theoretically propose and experimentally demonstrate a direct-detected optical OFDM transmission system with the assistance of a radio frequency (RF) tone. In the proposed system, an RF tone is inserted at the edge subcarrier of the OFDM signal to assist remote signal extraction after beating with all OFDM subcarriers at the receiver. This technique has been first proposed in [12] for direct-detected OFDM transmission. In this paper, we further show analytically that the proposed system can overcome the majority of the linear impairments from the transmitter to the receiver. Moreover, the carrier-to-signal-power ratio (CSPR) for both the previous SSB-OFDM and our proposed RF-tone-assisted OFDM systems are analytically analyzed. We observe that the relatively worse sensitivity of the previous SSB-OFDM systems can be attributed to the limited CSPR, resulted from the nonlinearity of the MZM. With this technique, the optimum CSPR for the best sensitivity can be easily achieved by controlling the relative amplitudes of the electrical input signal and the RF tone. We demonstrate a 10-Gb/s RF-tone-assisted OFDM system with an 8 quadrature-amplitude modulation (QAM) format. The experimental results show that the sensitivity of the proposed OFDM is ∼5 dB better than the previous intensity-modulated

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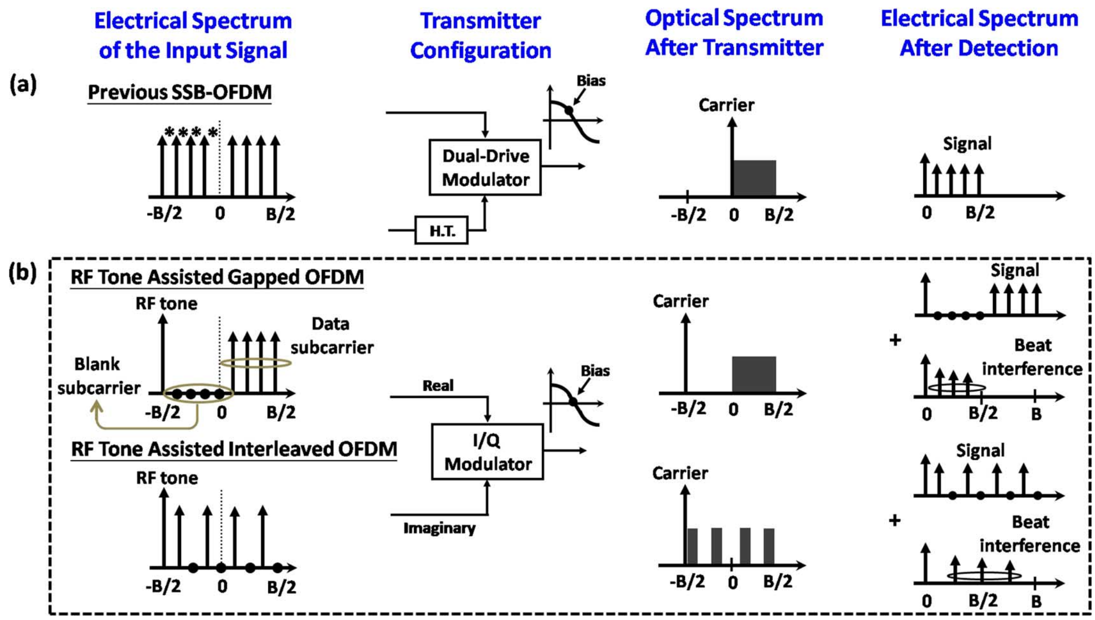

Fig. 1. Concepts for (a) the previous intensity-modulated SSB-OFDM, and (b) the proposed gapped and interleaved RF-tone assisted OFDM. H.T.: Hilbert transform.

SSB-OFDM systems. We also demonstrate 260-km uncompensated standard single-mode fiber (SSMF) transmission with negligible power penalty.

II. Principle of Operation

Fig. 1(a) shows the principle of the previous intensity-modulated SSB-OFDM [8], [9]. The data and its conjugate are converted to a time-domain real-valued electrical signal via the inverse fast Fourier transform (IFFT). This real-valued electrical signal and its Hilbert transform are fed into the two arms of a quadrature biased DD-MZM, resulting in an SSB-OFDM signal. The generation of the SSB can be understood more precisely through a mathematical modeling as follows. The discrete-time modulated optical signal can be written as

$$E(n) = \frac{1}{2}\left[\exp\left(j\frac{S(n)}{V_\pi}\pi\right) + j\exp\left(j\frac{\hat{S}(n)}{V_\pi}\pi\right)\right] \quad (1)$$

where $S(n)$ is the OFDM real-valued input with

$$S(n) = \sum_{k=1}^{N_d}\left[d_k \exp\left(j\frac{2\pi kn}{N}\right) + d_k^* \exp\left(-j\frac{2\pi kn}{N}\right)\right] \quad \text{(2-a)}$$

and $\hat{S}(n)$ stands for the Hilbert transform of $S(n)$, that is [13], [14]

$$\hat{S}(n) = \sum_{k=1}^{N_d}\left[\pm j d_k \exp\left(j\frac{2\pi kn}{N}\right) + (\pm j d_k)^* \exp\left(-j\frac{2\pi kn}{N}\right)\right] \quad \text{(2-b)}$$

where $j = \sqrt{-1}$, the asterisk means the complex conjugation, N_d and N are the numbers of the data and total subcarriers, n and k are the discrete-time index and the subcarrier index, respectively, d_k and d_k^* are the symbol and the conjugate of the symbol on the kth subcarrier, and V_π is the switching voltage of the MZM. From (2-b), we found that $\hat{S}(n)$ is also a real-valued signal and can be simply generated through rotating the modulated symbols d_k in $S(n)$ by $-90°$ or $+90°$ for upper or lower sideband modulation, respectively. Assuming that a small-signal input is used, which means $S(n), \hat{S}(n) \ll V_\pi$, then the optical signal could be simplified as

$$\begin{aligned} E(n) &\cong \frac{1}{2}(1+j) + j\frac{\pi}{2V_\pi}[S(n) + j\hat{S}(n)] \\ &= \frac{1}{2}(1+j) + j\frac{\pi}{V_\pi}\left[\sum_{k=1}^{N_d} d_k \exp\left(j\frac{2\pi kn}{N}\right)\right] \quad \text{(3-a)} \end{aligned}$$

for upper sideband modulation, or

$$E(n) = \frac{1}{2}(1+j) + j\frac{\pi}{V_\pi}\left[\sum_{k=1}^{N_d} d_k^* \exp\left(-j\frac{2\pi kn}{N}\right)\right] \quad \text{(3-b)}$$

for lower sideband modulation.

A perfect SSB-OFDM could only be achieved with relatively small optical modulation index (OMI), where $\mathrm{OMI} = (V_{\mathrm{in}})_{\mathrm{RMS}}/V_\pi$ with $(V_{\mathrm{in}})_{\mathrm{RMS}}$ the root-mean-square (RMS) amplitude of the electrical input to MZM. Note that the definition for OMI in this paper is different from that for the conventional OMI ($\mathrm{OMI_c}$), which is defined as the ratio of half the optical peak-to-peak intensity variation $(\Delta P)_{\mathrm{pp}}$ to the average optical power P_{avg}, that is, $\mathrm{OMI_c} = (\Delta P)_{\mathrm{pp}}/(2P_{\mathrm{avg}}) \propto (V_{\mathrm{in}})_{\mathrm{pp}}/V_\pi$, where $(V_{\mathrm{in}})_{\mathrm{pp}}$ is the peak-to-peak amplitude of the electrical signal to MZM. Since the OFDM signal has multiple subcarriers modulated by multiple independent data, its amplitude would behave like a complex Gaussian distribution. Thus, in this paper, the RMS amplitude, instead of the peak-to-peak amplitude, would be more suitable for evaluating OMI of the OFDM signal. A larger OMI typically used for higher modulation efficiency would destroy the small-signal assumption and introduce extra higher order nonlinear terms from the raised-cosine transfer function of the MZM. At the receiver, a photodiode converts the optical power into photocurrent directly since the signal is intensity modulated. Without introducing the signal–signal beat interference (SSBI), there is no frequency gap required between the signal and the carrier, and thus this SSB-OFDM is very spectrally efficient.

Fig. 1(b) shows the proposed two kinds of RF-tone-assisted OFDM systems with two possible subcarrier allocations. For convenience, we name the two systems as the gapped OFDM and the interleaved OFDM, respectively. For both systems, one RF tone for remote direct detection is inserted at the leftmost subcarrier. For the gapped OFDM, the blank subcarriers with no power, which are reserved for SSBI, are clusterly allocated between the inserted RF tone and the signal band. The number of the blank subcarriers is the same as that of the data subcarriers. The frequency gap formed by the blank subcarriers leads to the name of the gapped OFDM. For the interleaved OFDM, an RF tone is also used and placed at the left edge of the signal band. The blank and data subcarriers are now interleavedly allocated as shown in Fig. 1(b). Because of this interleaved manner, we call this scheme as the interleaved OFDM. For both approaches, the electrical OFDM signal after IFFT is complex and its real and imaginary parts are fed into the two arms of an optical I/Q modulator. Note that the bias point for the I/Q modulator is exactly set at the null to completely suppress the original optical carrier. The modulated optical spectra are shown after the I/Q modulators in Fig. 1(b). For both approaches, the optical spectra, including the carrier, are essentially equal to those of their electrical driven signal and have a bandwidth of B where B is the electrical bandwidth of the RF signal. Since the signal is linear-field modulated, the SSBI is inevitably introduced to the received signal after a square-law photodiode. For the gapped OFDM, the SSBI falls into the gap and does not affect the data subcarriers. As for the interleaved OFDM, the SSBI falls between the data subcarriers, therefore does not interfere with the data. Thus, the data of the two schemes can be extracted without being interfered by the SSBI.

Here are the mathematical models for the proposed OFDM schemes. The discrete-time baseband model for the RF-tone-assisted systems after the optical I/Q modulator can be represented as [15], [16]

$$E(n) = \sin\left(\frac{\pi}{V_\pi}\mathrm{Re}[S(n)]\right) + j\sin\left(\frac{\pi}{V_\pi}\mathrm{Im}[S(n)]\right) \cong KS(n) \quad (4)$$

where the operations of $\mathrm{Re}[x]$ and $\mathrm{Im}[x]$ give the real part and the imaginary part of x, respectively. $S(n)$ is the input OFDM signal with the inserted RF tone and can be expressed as

$$S(n) = A\exp(-j\pi n) + \sum_{k=0}^{N_d-1} d_k \exp\left(j\frac{2\pi f(k)n}{N}\right) \quad (5)$$

where N_d and N are the numbers of the data and total subcarriers with $N = 2N_d, n$ and k are the discrete-time index and the subcarrier index ranging from $(-N/2)$ to $(N/2-1)$, respectively, A is the given amplitude of the RF tone located at the $(-N/2)$th subcarrier, d_k stands for the data symbol on the kth subcarrier, and $f(k) = 0, 1, 2, \ldots, (N/2-1)$ for the gapped OFDM and $f(k) = (-N/2)+1, (-N/2)+3, \ldots, N/2-3, N/2-1$ for the interleaved OFDM. Equation (4) shows that $E(n) \cong KS(n)$ with $K = \pi/V_\pi$ when the input $S(n)$ is small enough. We assume that $E(n) = S(n)$ for simplifying the following analysis, which should be reasonable when a small modulation depth is utilized for the MZM. Naturally the assumption of $K = 1$ is also made without affecting the conclusions in this paper.

At the receiver, the photocurrent $I(n)$ can be modeled as the square of the absolute value of $E(n)$, i.e., $I(n) = R|E(n)|^2 = I_{\mathrm{dc}} + I_{\mathrm{SSBI}}(n) + I_{\mathrm{sig}}(n)$, where R is the responsivity of the photodiode, I_{dc} is the yielded direct current (dc), $I_{\mathrm{SSBI}}(n)$ stands for the SSBI, and $I_{\mathrm{sig}}(n)$ is the desired data current resulted from the beats between carrier and the data subcarriers, and can be written as

$$I_{\mathrm{sig}}(n) = RA^* \sum_{k=0}^{N_d-1} d_k \exp\left(j\frac{2\pi[f(k)+N/2]n}{N}\right) + c.c. \quad (6)$$

III. Channel Equalization

In this section, we will show how the proposed schemes can effectively compensate the majority of the linear channel impairments throughout the transmission link. We denote the discrete frequency responses of the transmitter, fiber CD, optical filters, and the receiver as $H_T(k), H_{\mathrm{CD}}(k), H_O(k), H_R(k)$, respectively. The fiber CD linearly distorts the optical phase on each subcarrier and can be simply modeled as $H_{\mathrm{CD}}(k) = \exp(j\pi DL(\lambda^2)/(c)(k\Delta f)^2)$ [17], where D is the dispersion parameter, L is the fiber length, λ is the operating wavelength, Δf is the frequency spacing of the subcarriers, and c is the speed of light in vacuum. The optical filter response $H_O(k)$ represents the overall effect of all the optical filters cascaded in line, i.e., $H_o(k) = \prod_{\forall i} H_{o,i}(k)$, where $H_{o,i}$ stands for the ith optical filter in the link. The overall frequency response before the photodiode could be modeled as the multiplication of the frequency response of the transmitter, fiber CD, and all the optical filters: $H_C = H_T H_{\mathrm{CD}} H_O$. Then, the discrete-time optical waveform before the photodiode $E_t(n)$ can be written as

$$E_t(n) = \begin{bmatrix} AH_C(-N/2)\exp(-j\pi n) \\ +\sum_{k=0}^{N_d-1} d_k H_C(f(k)) \exp\left(j\frac{2\pi f(k)n}{N}\right) \end{bmatrix}. \quad (7)$$

After the photodiode with a responsivity of R, the photocurrent can be expressed as

$$\begin{aligned} I(n) &= R|E_t(n)|^2 \\ &= \left\{ R\left[\sum_{k=0}^{N_d-1} d_k H_C(f(k)) A^* H_C^*(-N/2) \right.\right. \\ &\qquad \left.\left. \exp\left(\frac{j2\pi(f(k)+N/2)n}{N}\right)\right] + c.c.\right\} \\ &\quad + I_{\mathrm{SSBI}}(n) \end{aligned} \quad (8)$$

where the first term is the desired signal $I_{\rm sig}(n)$. Considering the discrete frequency response of the receiver $H_R(k)$, the desired signal becomes

$$I_{\rm sig}(n)=R\sum_{k=0}^{N_d-1}\left[\begin{Bmatrix} d_kH_C(f(k))H_R(f(k)+N/2) \\ \times A^*H_C^*(f(k))\exp\left(j\dfrac{2\pi(f(k)+N/2)n}{N}\right)\end{Bmatrix}\right] + c.c. \quad (9)$$

After taking the fast Fourier transform (FFT), the extracted symbol on each subcarrier with distortion is

$$R_k=R[d_kH_C(f(k))H_R(f(k)+N/2)A^*H_C^*(-N/2)]. \quad (10)$$

The training symbols at the start of each OFDM packet will compare the received symbol R_k and the known symbols d_k at the receiver, and derive the channel model of $H(k)=R_k/d_k=RH_C(f(k))H_R(f(k)+N/2)A^*H_C^*(-N/2)$. The transmitted data following the training sequence will be equalized by multiplying the inverse of $H(k)$ to the received symbol, i.e., $D_k=R_k/H(k)$, and thus all the linear impacts from the transmitter, fiber CD, cascaded optical filtering, and the receiver can be compensated theoretically. This simple equalization technique is called the one-tap equalizer or zero-forcing equalizer with the disadvantage of enhancing the noise for those severely attenuated subcarriers [18]. However, considering an optical transmission system in which the PMD-induced fading is not so significant, i.e., the data subcarriers are not strongly attenuated by the PMD fading, the zero forcing is still the best choice due to its simplicity. Due to the above reasons, we choose this equalization technique throughout this paper.

Note that throughout the whole analysis we assume the OMI to be small for ensuring the signal is linearly modulated and not interfered by the nonlinear distortions from the MZM. However, a small OMI would cause significant excess modulation loss at the transmitter. The similar tradeoff between the power loss and the MZM nonlinearity has been discussed previously, and can be solved by employing the electronic predistortion [16].

IV. Carrier-to-Signal-Power Ratio

In this section, we will derive the relation between the CSPR and received electrical signal power for both the previous intensity-modulated SSB-OFDM and proposed schemes.

For RF-tone-assisted OFDM, the average optical power $P_{\rm avg}$ for an optical field of $E(n)$ can be written as the summation of the RF tone power and the signal power, i.e., $P_{\rm avg}=\varepsilon[|E(n)|^2]=|A|^2+N_d\varepsilon[|d_k|^2]$, where $\varepsilon[x]$ stands for the expectation of x. The definition of CSPR is expressed as $\mathrm{CSPR}=|A|^2/N_d\varepsilon[|d_k|^2]$. The optimum CSPR is defined as the power ratio between the RF tone and the signal to gain the maximum detected electrical power for a fixed optical received power $P_{\rm avg}$.

After the photodiode, the electrical power $P_{\rm el}$ of the desired signal photocurrent $I_{\rm sig}(n)$ can be expressed as

$$\begin{aligned}P_{\rm el}&=\varepsilon[|I_{\rm sig}(n)|^2]=2R^2A^2N_d\varepsilon[|d_k|^2]\\ &\le\frac{1}{2}R^2(A^2+N_d\varepsilon[|d_k|^2])^2. \end{aligned}\quad (11)$$

For a fixed received optical power $P_{\rm avg}$, (11) gives the upper bound of $1/2R^2(A^2+N_d\varepsilon[|d_k|^2])^2$ for the electrical power $P_{\rm el}$, and this bound is achieved when the RF-tone power equals the signal power, i.e., $A^2=N_d\varepsilon[|d_i|^2]$. Thus, the optimum CSPR, which notably is obtained independently to the number of the data subcarriers or the QAM size, is found to be 0 dB and matches the previous numerical results depicted in [19]. With this optimum CSPR, the received electrical power can be related to the input optical power as $P_{\rm el}=R^2P_{\rm avg}^2/2$. Note that the above derivation only considers the back-to-back (b2b) case while it can be easily derived in the transmission case by incorporating the channel response H_C into considerations in the above analysis.

Now we derive the CSPR for the conventional intensity-modulated SSB-OFDM. To simplify the analysis, we assume the transfer function of the MZM to be linear. The received continuous photocurrent $I_{\rm sig}(t)$ and the average electrical signal power $P_{\rm el}$ can be written as a function of the received optical power $P(t)$, the average optical power $P_{\rm avg}$, and OMI as follows:

$$I_{\rm sig}(t)=RP(t)=2RP_{\rm avg}P(t)/(2P_{\rm avg}) \quad (12)$$

$$P_{\rm el}=\varepsilon\left[I_{\rm sig}^2(t)\right]=4R^2P_{\rm avg}^2\mathrm{OMI}^2. \quad (13)$$

It is shown that the received electrical power is proportional to the square of OMI and thus can be enhanced by directly increasing the OMI at the transmitter. If we define the peak-to-average-power ratio (PAPR) of the input OFDM signal in the transmitter as $\mathrm{PAPR}=V_{\rm peak}^2/V_{\rm RMS}^2=V_{\rm peak}^2/(\mathrm{OMI}^2V_\pi^2)$ where $V_{\rm peak}$ is the peak amplitude of the OFDM input, then the electrical power can be related to the PAPR as $P_{\rm el}=4R^2P_{\rm avg}^2V_\pi^2/(V_{\rm peak}^2\mathrm{PAPR})$. If the peak value of the input of voltage $V_{\rm peak}$ reaches the maximum and minimum of the transfer function of the MZM, i.e., $2V_{\rm peak}=V_\pi$, then

$$P_{\rm el}=R^2P_{\rm avg}^2/\mathrm{PAPR}. \quad (14)$$

The PAPR of an OFDM signal is typically high due to the Gaussian distributed waveform, and therefore, the received power would be several times smaller than $R^2P_{\rm avg}^2/2$, which is the received power of our proposed systems. Clipping is one possible solution to reduce the PAPR for the previous SSB-OFDM systems, i.e., $\mathrm{PAPR}=2$ would result in the same received electrical power as our proposed system. However, the clipping will introduce extra distortion and degrade the receiving performance. Naturally, our proposals would exhibit a better sensitivity compared to the intensity-modulated SSB-OFDM.

V. Experimental Setup

Fig. 2 shows the experimental setup of the proposed RF-tone-assisted OFDM systems. The required digital signal processing (DSP) blocks for the OFDM signals are emulated and prerun by the Matlab program. Ninety two subcarriers for the data and one subcarrier for the RF tone are zero padded with a size of 256 and converted to the time-domain waveform through the use of IFFT. The data rate is 10 Gb/s, which is mapped onto a circular 8-QAM and thus the data rate of each subcarrier is 36 M symbols per second. The duration of an OFDM symbol is

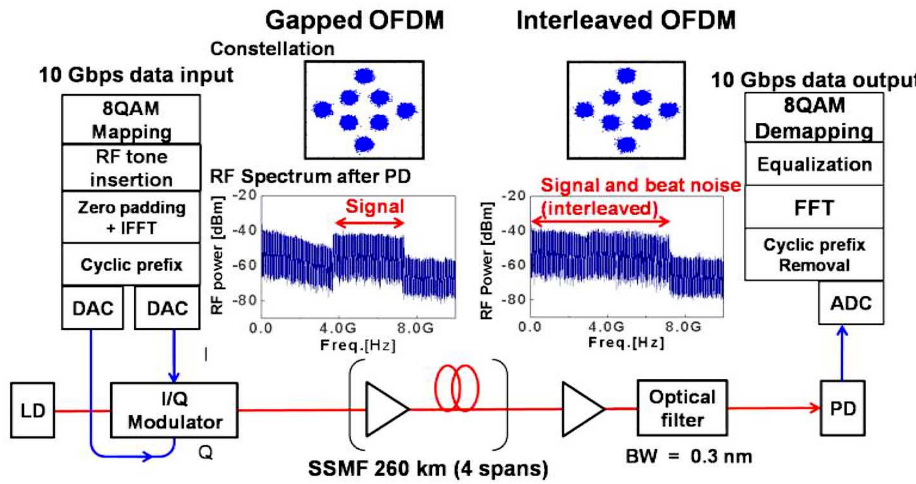

Fig. 2. Experimental setup of the RF-tone-assisted OFDM systems.

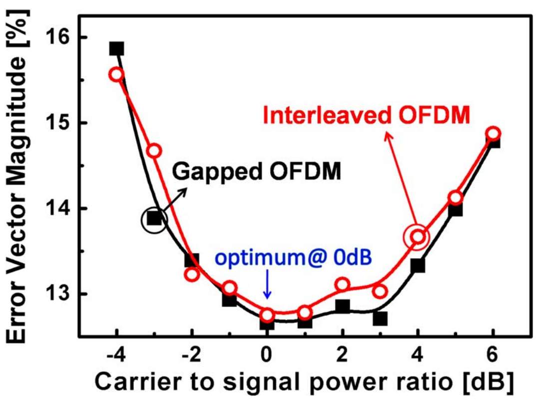

Fig. 3. Experimental results of error vector magnitude versus the CSPR.

27.6 ns and the cyclic prefix is 1.6 ns, which occupies $\sim 1/17$ of one OFDM symbol. Although we have used 30 training symbols preceding the transmitted 330 OFDM symbols for channel estimation, the net data rate could still be claimed as ~10 Gb/s because the training overhead typically is negligible for direct-detected OFDM as tested under the laboratory conditions [1]. The OFDM signal generated by Matlab is then loaded into an arbitrary waveform generator (Tektronix, AWG7102) with a sampling rate of 10 GHz. The two outputs of the arbitrary waveform generator (AWG), which are the real and imaginary parts of the OFDM signal, are low-pass filtered to remove the out of band image noise and then continuously sent into the two arms of an optical I/Q modulator, respectively. The pre-emphasis technique is used at the transmitter to achieve equal power output for all data subcarriers [20]. The output of I/Q modulator is sent into four spans of SSMF with a total distance of 260 km without optical dispersion compensation. Since the transmission link is setup by cascading multiple fiber spans (four spans), the fiber distance 260 km is currently limited by the available fiber spans and optical amplifiers [Erbium-doped fiber amplifiers (EDFA)] in the lab. At the receiver, the received signal is amplified and filtered with a 0.3-nm bandpass filter before being detected by a 10-GHz photodiode. The photocurrent is then sampled at 20 GHz and stored in a real time scope (Tektronix, TDS6604). The stored samples are postprocessed in Matlab for synchronization, cyclic prefix removal, FFT, and equalizations.

The insets shown in Fig. 2 are the equalized 8-QAM constellations with $\sim 30\,000$ symbol points and the corresponding RF spectra after the photodiode for both the gapped and interleaved systems. For the RF spectra, the signal band of the gapped OFDM is separated from the SSBI due to the use of frequency gap, while for the interleaved OFDM, the signal and the SSBI are interleaved after the photodiode. For both schemes, the occupied signal bandwidths are equal and close to 7.6 GHz for an 8-QAM, 10-Gb/s data rate.

VI. Experimental Results

Fig. 3 depicts the error vector magnitude (EVM) versus the CSPR. The EVM is used here to quantify the signal quality and a lower EVM value typically means a better performance.

Its rigorous definition is given as [21]

$$\mathrm{EVM}[\%] = 100 \left[\sum_{i=1}^{N_s} |R_i - d_i|^2 / N_s\right]^{1/2} \Big/ |d_{\max}|$$

where N_s is the total number of the transmitted symbols, d_i and R_i are the transmitted and received symbols, respectively, and $d_{\max}$ represents the symbol point with the maximum amplitude in the constellation diagram. For both the gapped and interleaved OFDM, the optimum CSPR is found to be ~0 dB, which implies that the best sensitivity is achieved when the carrier power equals to the signal power. These measured results match the above analytical derivation in Section IV and the numerical results studied in [19]. Note that pre-emphasis [20] to compensate for the limited bandwidth of AWG, electrical drivers, and filters is implemented to ensure that the output optical CSPR matches the prespecified CSPR in Matlab.

In Fig. 4, we compare the nonlinear tolerance for both the gapped and interleaved OFDM systems by varying the optical input power to each fiber span. After transmitted 260 km of uncompensated SSMF, we found that the interleaved OFDM is more robust to the fiber nonlinearities possibly due to the larger subcarrier spacing which can suppress the four wave mixing (FWM). The signal constellations shown in the insets of Fig. 4,

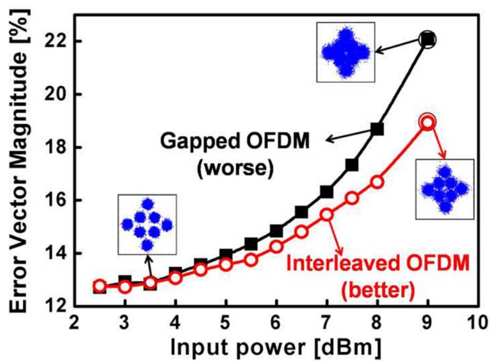

Fig. 4. Experimental results of error vector magnitude versus the input power per span.

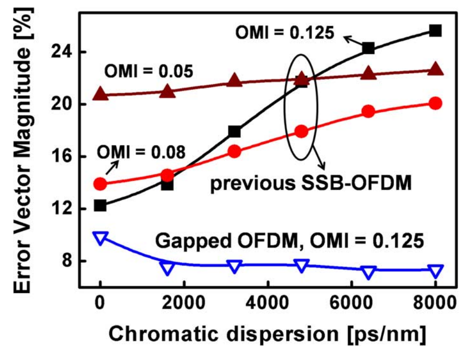

Fig. 6. Simulation of the EVM as a function of the chromatic dispersion for the previous intensity-modulated SSB-OFDM and our proposed RF-tone-assisted gapped OFDM.

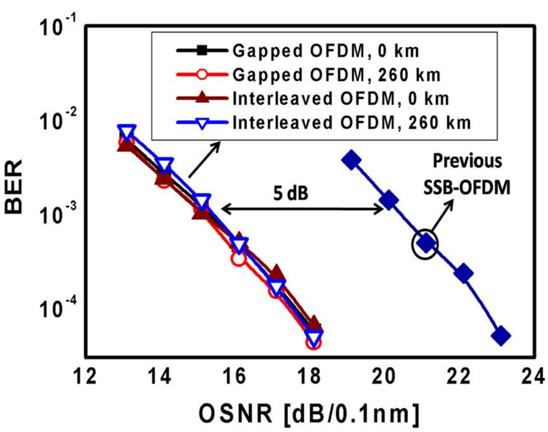

Fig. 5. Experimental results of BER versus OSNR.

with the same received OSNR, indicate better signal quality of the interleaved OFDM with higher optical input power.

In Fig. 5, the bit error rate (BER) performance is implemented for both the proposed and the previous intensity modulated SSB-OFDM system at 10 Gb/s with 8 QAM. The BER is evaluated by counting the number of different bits between the transmitter and receiver, and therefore, its range is constrained high (typically $> 10^{-5}$) due to the limited amount of the testing data. The CSPR is fixed at 0 dB for our proposed schemes. The optimized OMI $\sim = 0.12$ in b2b is used for the previous SSB-OFDM. The b2b sensitivities of the gapped and interleaved OFDM systems are very close and both systems require an optical signal-to-noise ratio (OSNR) of $\sim$15.2 dB with 0.1-nm optical resolution bandwidth to achieve a BER of 10^{-3}. The previous SSB-OFDM system is found with an $\sim$5-dB worse sensitivity than our proposed systems because of the constrained CSPR limited by the nonlinearity of the MZM. There is negligible penalty observed for our schemes after 260-km SSMF transmission. While for the previous SSB-OFDM system with the optimized b2b OMI, the BER could only reach 2.3×10^{-3} at an OSNR $= 30.3$ dB.

Fig. 6 shows the numerical results of CD tolerances for the previous SSB-OFDM and the proposed gapped OFDM schemes considering the nonlinearity of the MZM. For both systems, no clipping is used and the fiber CD is modeled by the frequency response $H_{\rm CD}(k)$ described in Section III. For both schemes, the data rate is 10 Gb/s with 8 QAM. The optimized OMI for the SSB-OFDM in b2b is found to be $\sim$0.125, which has driven the MZM into the nonlinear region. To make a fair comparison, the used OMI for our gapped OFDM is also $\sim$0.125 but with the OMI definition of $(V_{\rm in})_{\rm RMS}/(2V_\pi)$. Compared to the OMI of the previous system, the extra factor of two in the denominator is resulted from the doubled maximum voltage swing biasing at the null. From the figure, we can see the tradeoff between the sensitivity and the CD tolerance in previous SSB-OFDM. This tradeoff can be explained as follows: when a small OMI is used with a poor sensitivity based on (13), this intensity-modulated signal will behave like a field-modulated signal and thus has a better CD tolerance similar to our proposed system. On the other hand, when a larger OMI is employed for a better sensitivity, the CD, which linearly distorts the optical field, would turn to a nonlinear distortion after the photodiode, therefore diminishing the efficiency of the linear zero-forcing equalizer and reducing the CD tolerance. Since for our system the sensitivity can be optimized independently of the OMI and the signal is field modulated, the b2b performance and the CD tolerance are both improved compared to the previous SSB-OFDM, which validates our experimental results.

The proposed direct-detected OFDM has a $\sim$7-dB worse sensitivity when compared with the 10-Gb/s coherent OFDM, for which the required OSNR at BER $= 10^{-3}$ is $\sim$8 dB after scaling [22]. This $\sim$7-dB OSNR penalty could be attributed to the inherent optical carrier ($\sim$3 dB when CSPR $= 0$ dB) and the “subcarriers X ASE” noise (up to 3–5 dB depending on the bandwidth of the optical filter [23]) in the direct-detected OFDM. However, both the sensitivity and spectra efficiency of the direct-detected OFDM could be possibly improved by boosting the carrier power relative to the sideband just prior to photodetection [24].

VII. Conclusion

We have theoretically analyzed and experimentally demonstrated the proposed RF-tone-assisted OFDM with two different schemes of the gapped and interleaved OFDM. We show via analytical results that the proposed systems could equalize almost all the linear impairments from the link, and we also analytically obtain the optimum CSPR to be 0 dB for the proposed system. We transmit 10 Gb/s data with a QAM size of 8 through 260-km uncompensated SSMF. The experimental results show that the gapped and interleaved OFDM outperform the previous SSB-OFDM by $\sim$5 dB b2b and with negligible penalty after 260-km SSMF transmission.

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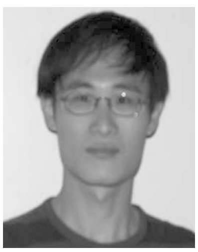

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