Microwave Photonic Direct Sequence Transmitter and Heterodyne Correlation Receiver

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Abstract **We propose and demonstrate a new architecture for the transmission, heterodyne reception, and correlation of direct sequence signals encoded onto an optical carrier. The approach is practical for pseudo-noise modulated laser radar and free space optical code division multiple access communication. Although the local oscillator is free running, we show that the received signal is free from laser phase noise. Furthermore, by applying direct sequence coding to both the transmitted signal and the local oscillator, the delay required for correlation can be realized via a combination of electrical and optical means.**

Index Terms **Laser Radar, Optical Modulation, Spread Spectrum Radar.**

I. Introduction

We present a spread-spectrum sub-carrier multiplexed (SCM) transmitter and a complimentary heterodyne correlation receiver. The architecture is applicable to both CW laser radar and optical code division multiple access (OCDMA) systems. This work focuses on the laser radar application.

The 1550-nm wavelength range is currently being explored for use in laser radar [1], [2] and free-space optical communications [3]. Eye-safety, the availability of inexpensive and reliable off-the-shelf telecommunications components, and wavelength division multiplexing capability are all compelling reasons to consider this wavelength for free space applications.

Range finding with both laser and microwave radar can be performed with continuous signals using waveforms that are frequency modulated or pseudo-randomly modulated [4]. When a pseudo-random bit sequence is used to perform the modulation, the resulting waveform is known as a direct sequence (DS) [5] spread spectrum signal.

CW laser radar with pseudo-random modulation and direct detection has been reported [6], [7]. However, un-amplified direct detection of small powers is impractical at 1550 nm, due to the lack of good avalanche photodiodes at this wavelength. Heterodyne detection is one straightforward means of improving sensitivity without the complexity, power consumption, and optical noise figure penalty of optical pre-amplification.

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Whereas DS encoding could potentially be applied directly to an optical carrier, there are many potential benefits to using a DS encoded microwave sub-carrier. Foremost for the laser radar application is the ability to use the microwave sub-carrier to perform cancellation of the optical phase noise [8]. This allows even noisy distributed feedback lasers (DFBs) to be used for heterodyne detection. Furthermore, direct sequence OCDMA systems with a microwave sub-carrier [9] can potentially benefit from full code orthogonality and a reduction in cumulative shot noise.

In our architecture we combine the benefits of the spread spectrum waveform with heterodyne detection, sub-carrier multiplexing, and phase noise cancellation. Furthermore, we use photonic techniques for both the generation and correlation of the spread spectrum waveform.

II. Principle of Operation

A. System Layout

A block diagram of our direct sequence transmitter and receiver is shown in Fig. 1. The transmitted optical signal is encoded with a pseudorandom digital sequence. The optical local oscillator (LO) used for heterodyne detection is likewise modulated with a delayed version of the same sequence. The two sequences are then multiplied together in the receiver to produce a correlation signal. We show that the amplitude of the correlation signal is negligible unless the time delay of the sequence matches the trip time from transmitter to receiver. This enables a range-gating function to be performed for laser radar applications. The transmitter and receiver are described in detail below.

B. Direct Sequence Transmitter

In the transmitter, light from a CW distributed feedback (DFB) laser is intensity modulated in a Mach Zehnder modulator (MZM) by a microwave tone at a frequency of ω_{RFt}. Following the analysis of Corral et al. [10] (and noting that our standard MZM is a particular case of the dual-drive modulator), the optical field following the modulator can be represented as

$$E_t(t) \propto \begin{Bmatrix} \sqrt{P_t}\exp[j(\omega_t t + \phi_t(t) + \phi_i)] \\ \cdot \sum_{n=-\infty}^{\infty} J_n(m)\cdot\cos[(\phi_{bt} - n\pi)/2]\cdot e^{jn(\omega_{RFt}t+\pi/2)} \end{Bmatrix} + c.c, \quad (1)$$

where, $J_n(x)$ is a Bessel function of the first kind of order n. P_t, ω_t, and ϕ_t(t)are the power, frequency, and phase of the transmitter s laser. ϕ_{bt}, and m, are the bias port controlled differential phase shift, and the optical modulation index. ϕ_i, is the insertion phase shift through the modulator when no voltages are applied. The modulated signal spectrum is illustrated in Fig. 2(a).

Furthermore, an N bit-long pseudo random binary bit sequence, a_i, is applied to the bias port of the MZM. The amplitude and offset voltage of the sequence is adjusted such that the modulator s bias port voltage shifts between $V_\pi/2$ and $3V_\pi/2$ when a bit changes state. This implies that ϕ_{bt} shifts between π/2 and 3π/2.

The transmitter output is then sent through a fiber delay line, attenuator, and polarization compensator. This simulates the delay and loss of free space transmission. Since we are using only 0.7 km of single mode fiber, we will assume that chromatic dispersion and the dispersion penalty [11] is small enough to be neglected.

In order to present a concise mathematical analysis, we also make the reasonable assumption that the optical modulation index, m, is less than unity. In this case the higher order terms of the series expansion in (1) rapidly approach zero. Retaining only the $n = 0$ and $n = -1$ terms in (1), the transmitted signal after propagation to the receiver s photodiode will be given by

$$E_t(t) \propto \sqrt{P_t}\cos\left[\omega_t\cdot(t-\tau_t) + \phi_t(t-\tau_t) + \phi_i\right] \cdot \begin{Bmatrix} J_0(m)\cdot a(t-\tau_t) \\ -2J_1(m)\cdot\sin[\omega_{RFt}\cdot(t-\tau_t)] \end{Bmatrix}, \quad (2)$$

where

$$a(t) = -1 + 2\cdot\sum_{i=0}^{N} a_i \varepsilon(t - iT), \quad (3)$$

$$\varepsilon(x) = \begin{Bmatrix} 1 \text{ for } (0 \le x < T) \\ 0 \text{ otherwise} \end{Bmatrix}, \quad (4)$$

T is the bit period, and τ_t is the optical path length delay between the transmitter s MZM and the receiver s photodiode.

Note that if this signal in (2) were squared, corresponding to direct detection of intensity in a photodiode, the microwave subcarrier at ω_{RFt} would be multiplied by the bipolar spreading sequence, a. As previously demonstrated [12], this is one method of generating BPSK modulation without using a RF mixer.

C. Local Oscillator

A free-running LO laser tuned to an optical frequency of ω_{lo} is used for heterodyne detection. The LO signal is externally modulated in a MZM using a microwave tone at a frequency of ω_{RFlo}. The spectrum of the LO signal is illustrated in Fig. 2(b). The digital sequence is first delayed by τ_{el} before being applied to the bias port of the receiver s MZM. As in the transmitter, the drive level is adjusted to shift the bias phase between π/2 and 3π/2 when a bit changes state.

After passing through a short length of fiber, the modulated LO signal is combined with the received signal through the 90% port of a 10%/90% coupler and sent to a photodiode. The modulated LO signal after propagation to the photodiode is then given by

$$E_{lo}(t) \propto \sqrt{P_{lo}}\cos\left[\omega_{lo}\cdot(t-\tau_{lo}) + \phi_{lo}(t-\tau_{lo}) + \phi_i\right] \cdot \begin{Bmatrix} J_0(m)\cdot a(t-\tau_{el}-\tau_{lo}) \\ -2J_1(m)\cdot\sin[\omega_{RFlo}\cdot(t-\tau_{lo})] \end{Bmatrix}, \quad (5)$$

where P_{lo}, ω_{lo}, and $\phi_{lo}(t)$ are the power, frequency, and phase of the LO laser s signal respectively, and τ_{lo} is the optical path length delay between the local oscillator s MZM and the photodiode. The modulation index, m, and insertion phase shift, ϕ_i, are assumed to be the same as in the transmitter.

D. Correlation Receiver

The current detected at the photodiode will consist of a homodyne and a heterodyne signal. The heterodyne signal, I_h, is given by

$$I_h(t) \propto \sqrt{P_{lo}P_t}\cos[\omega_b t - \omega_t\tau_t + \omega_{lo}\tau_{lo} + \phi_t(t-\tau_t) - \phi_{lo}(t-\tau_{lo})] \cdot \begin{Bmatrix} J_0^2(m)\cdot a(t-\tau_{el}-\tau_{lo})a(t-\tau_t) \\ -2\cdot J_0(m)J_1(m)a(t-\tau_{el}-\tau_{lo})\sin[\omega_{RFt}(t-\tau_t)] \\ -2\cdot J_0(m)J_1(m)a(t-\tau_t)\sin[\omega_{RFlo}(t-\tau_{lo})] \\ +4\cdot J_1^2(m)\sin[\omega_{RFt}(t-\tau_t)]\sin[\omega_{RFlo}(t-\tau_{lo})] \end{Bmatrix}, \quad (6)$$

where $\omega_b = \omega_t - \omega_{lo}$,is the beat frequency. As evident in (6) and illustrated in Fig. 2(c), the photodiode signal contains the beat frequency and several sidebands corresponding to the transmitter and receiver RF modulation frequencies.

Observe that (6) contains the laser phase noise terms, $\phi_t(t)$ and $\phi_{lo}(t)$ However, we now square the signal represented by (6), by sending the photodiode signal through a microwave square law detector. The squared heterodyne signal is given by,

$$I_h^2(t) \propto P_{LO}P_T A(t)\{1 + \cos[2\omega_b t - 2\omega_t\tau_t + 2\omega_{lo}\tau_{lo} + 2\phi_t(t-\tau_t) - 2\phi_{lo}(t-\tau_{lo})]\} \quad (7)$$

where

$$
\begin{aligned}
A(t) = & 8J_0^2(m)J_1^2(m)a(t-\tau_{el}-\tau_{lo})a(t-\tau_t)f_t(t)f_{lo}(t) \\
& +16J_1^4(m)f_t^2(t)f_{lo}^2(t)+J_0^4(m)+4J_0^2(m)J_1^2(m)f_t^2(t) \\
& +4J_0^2(m)J_1^2(m)f_{lo}^2(t)-2J_0^3(m)J_1(m)a(t-\tau_t)f_t(t) \\
& -2J_0^3(m)J_1(m)a(t-\tau_{el}-\tau_{lo})f_{lo}(t) \\
& -8J_0(m)J_1^3(m)a(t-\tau_t)f_{lo}^2(t)f_t \\
& -8J_0(m)J_1^3(m)a(t-\tau_{el}-\tau_{lo})f_t^2(t)f_{lo}(t)
\end{aligned}
\qquad (8)
$$

$$f_t(t)= \sin[\omega_{RFt}(t\text{-}\tau_t)], \qquad (9)$$

$$f_{lo}(t)= \sin[\omega_{RFlo}(t\text{-}\tau_{lo})]. \qquad (10)$$

The lowest frequency components from the output of the square law detector are illustrated in Fig. 2(d). The first line of (8) gives a signal component at the IF frequency, ω_{if} = ω_{RFt}— ω_{RFlo}. If we filter the output of the square law detectorto pass only the IF frequency, the filtered output, I_{sq}, will be

$$I_{sq}(t) \propto A_{if}(t)\cdot\cos(\omega_{if}t-\omega_{RFt}\tau_t+\omega_{RFlo}\tau_{lo}) \qquad (11)$$

where

$$A_{if}(t) = J_0^2(m)J_1^2(m)a(t-\tau_{el}-\tau_{lo})a(t-\tau_t) \quad \cdot \qquad (12)$$

Note that that this IF frequency component does not contain the optical phase noise terms, $\phi_t(t)$ and $\phi_{lo}(t)$. Kuri [8] demonstrated a similar method of optical phase noise cancellation by heterodyne detection with a dual mode optical local oscillator and a microwave square law detector.

In order to improve the signal to noise ratio, a band pass filter centered near ω_b may be inserted between the photodiode and the square law detector as shown in Fig. 1 and Fig. 2 (c). The filter would reject DC, the LO homodyne signal at ω_{RFlo} and the heterodyne frequency components $\omega_b\pm(\omega_{RFlo+}\omega_{RFt})$. Following the same analysis as above, but with the filtered components removed from the heterodyne signal, we find that the IF frequency signal is still given by (11) and (12).

Clearly, there are other valid filtering arrangements. For example, the bandpass filter could be set to pass only ω_b and the nearest two side-bands at $\omega_b\pm\omega_{if}$. In this case the IF frequency signal, although reduced in amplitude, would still be obtained.

It is also evident from (12) that the amplitude of the IF signal is proportional to the product of the transmitted sequence and its time-shifted replica. The time averaged amplitude, $f(t)$, of the IF signal is

$$
\begin{aligned}
f(\tau) & \equiv \lim_{T\to\infty}\frac{1}{2T}\int_{-T}^{T}A_{if}(t)\cdot dt \\
& \propto \lim_{T\to\infty}\frac{1}{2T}\int_{-T}^{T}a(t-\tau_{el}-\tau_{lo})\cdot a(t-\tau_t)\cdot dt\,, \qquad (13) \\
& \propto \int_{-\infty}^{\infty}a(t'-\tau)\cdot a(t')\cdot dt'
\end{aligned}
$$

where,

$$\tau = \tau_{el}+\tau_{lo}-\tau_t\,. \qquad (14)$$

Note that $f(\tau)$ is also the autocorrelation function of the pseudorandom signal, $a(t)$. Thus, if $a(t)$ is a bipolar maximal length shift register sequence (*M*-sequence), then $f(\tau)$ will be nearly zero unless $|\tau| < T$. This makes range-gated measurement in the laser radar application straightforward.

Also observe that the phase of the IF signal in (11) depends upon $\omega_{RFt}\tau_t$. This implies that an approaching or receding target will produce phase modulation. Hence, with some additional processing, the target velocity could be determined.

III. Experiment

We tested the system using the configuration described above. The following frequencies were carefully chosen by considering component bandwidths and the location of spurious frequencies with respect to the desired signal: ω_{RFt} = (2π)Æ5 GHz, ω_{RFlo} = (2π)Æ3.5 GHz, and ω_b⁻ (2π)Æ11.6 GHz.

The digital sequence, a_i, was an N=2^{12}-1 bit long maximal length sequence (*M*-sequence) with a bit period of T = 36.6 ns.

In order to verify the bias settings of the transmitter s MZM, the transmitted DS signal was directly detected (no heterodyne) at the photodiode. The measured spectrum, shown in Fig. 3, was centered at 5 GHz and contained nulls spaced by $1/T$ (27.3 MHz). This was as expected for BPSK modulation. The operation of the receiver s MZM was verified in the same manner.

In order to evaluate the heterodyne performance, both lasers were operated with the microwave sub-carriers turned on, but the DS coding turned off (all 0 s). As shown in Fig. 4, the line-width of the measured heterodyne signal was broad. In the absence of the DS coding, the observed line-width was due to the (approximately 1 MHz) line-width of the DFB laser and the (approximately 100 kHz) linewidth of the tuneable laser.

The transmitter-to-receiver electrical delay of the code, τ_{el}, was then set to approximately 100 bits or 3.66 s. This matched the transmitter-receiver delay due to propagation through 0.75 km of single mode fiber.

With the DS code and both microwave carriers applied, the electrical delay was fine-tuned until the IF signal from the square law detector output was maximized. This corresponded to the ideal condition of $\tau = 0$.

The observed IF signal is shown in Fig. 5. Compare the narrow line-width of this signal to the broad line-width of the heterodyne signal shown in Fig. 4. As the theoretical analysis above predicts and this measurement verifies, the laser phase noise was indeed cancelled.

Furthermore, as the relative delay between the transmitter and receiver codes was changed, the carrier amplitude decreased according to the autocorrelation function of the *M*-sequence in excellent agreement with (13). The relationship is shown in Fig. 6.

In this demonstration, the received power level (measured at the photodiode input) was attenuated to 33 W. For comparison, consider that > 4 mW of received power would be required to reach shot noise limited direct detection with this photodiode. The benefit of heterodyne detection is evident. Nevertheless, sensitivity could be improved by replacing the square law detector used here with an optimally filtered product detector [8], [13].

IV. Discussion

This architecture offers a number of unique features. First, it should be obvious that the pseudo-random sequence could be removed from the LO laser and applied instead to a mixer at the IF stage. However, by applying the sequence to both the transmitter and the LO through an MZM, it is possible to generate the required correlation delay by using a combination of electrical delay and switched optical [14] delay. This would be particularly relevant if we chose to use a sequence that was non-deterministic such that it could not be replicated by use of a formula. In such a case, the required delay could be obtained optically.

Furthermore, the BPSK codes are applied optically rather than through a mixer. Hence, the RF carrier frequencies and code rate can be chosen from anywhere within the MZM s operational bandwidth, typically DC to greater than 10 GHz. A good balanced mixer generally has a much smaller range of operational frequencies.

The approach was also shown to be immune to laser phase noise. Since our modulation is imposed on a microwave sub-carrier, the receiver is designed to be sensitive to the sub-carrier phase rather than the optical phase. Hence, a simple square law detector was shown to be an effective means of canceling the optical phase noise. However, if filters are used before the square law or product detector, they must exhibit sufficient bandwidth to account for the laser line-width.

One important advantage of this approach is that the bipolar nature of the pseudorandom sequence is fully preserved throughout the system. By comparison, the cross-correlation, autocorrelation, and inter-code correlation properties of bipolar codes are not generally preserved in conventional direct detection optical systems. This serious limitation has led to the development of alternative sparse optical orthogonal codes [15], codes that operate with two wavelengths simultaneously [16], and other esoteric sequences [17]. We do not require any of these alternative sequences; rather, bipolar sequences that have been developed for wireless code division multiple access (CDMA) and RF radar ranging can be used without modification.

While coherent detection systems are generally phase preserving, and therefore compatible with bipolar codes, these systems generally require either exceptionally low laser phase noise or a complex optical phase locked loop setup.

Just as in a conventional CDMA wireless system, a replica of the transmitted code can be stored locally and synchronized in the receiver. Therefore, the transmitter and receiver do not need to be co-located. Hence, many of these techniques are directly applicable to OCDMA communication as well.

It is important to note that the system was built entirely with off-the-shelf 1550-nm telecommunications components, and that the 1550-nm wavelength is also advantageous for free space operation when eye-safety is a concern.

Finally, we realize that we have not discussed potential problems due to mode and polarization mismatch. However, these problems are not specific to this architecture, but are common to any heterodyne detection system. One solution is the polarization diversity receiver [18], whereby two receivers operate in parallel, one for each orthogonal polarization state.

V. Conclusion

In conclusion, we have demonstrated a novel architecture for the generation and correlated reception of direct sequence coded optical signals. The approach is relevant to both laser radar and OCDMA.

We have used DS coding in order to obtain range resolution from a CW signal. Furthermore, by using a microwave sub-carrier and a microwave product detector, we are able to cancel optical phase noise from the received correlation signal. In addition, the DS code is applied directly to the MZM rather than through a microwave mixer. Since the DS signal is also modulated onto the receiver s LO laser, the delay required for correlation can be realized by a combination of electrical and optical delay.

To our knowledge this is the first demonstration and analysis of a heterodyne reception architecture that combines the features of bipolar DS coding and optical phase noise cancellation. We expect that this will lead to the development of practical, inexpensive, laser radar and free space OCDMA systems built with off-the-shelf telecommunications components.

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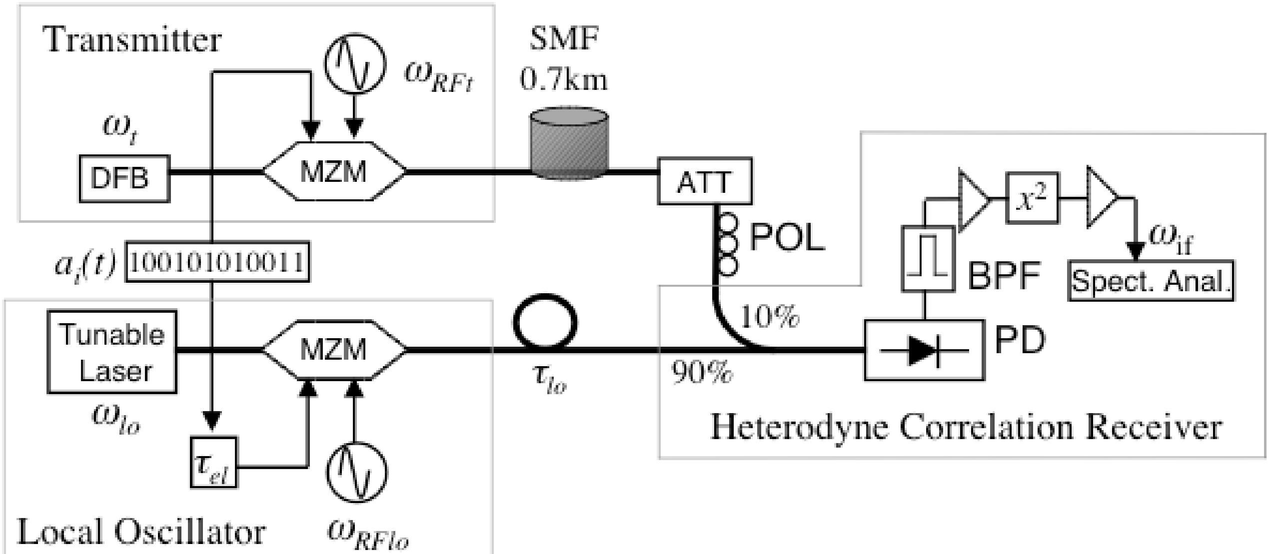

Fig. 1. Block diagram of Transmitter, Local Oscillator, and Correlation Receiver. MZM:Mach Zehnder Modulator, DFB: Distributed Feedback Laser, SMF single mode fiber, ATT: attenuator, POL: polarization controller, PD: photodiode, BPF:band-pass filter. Variables are explained in text.

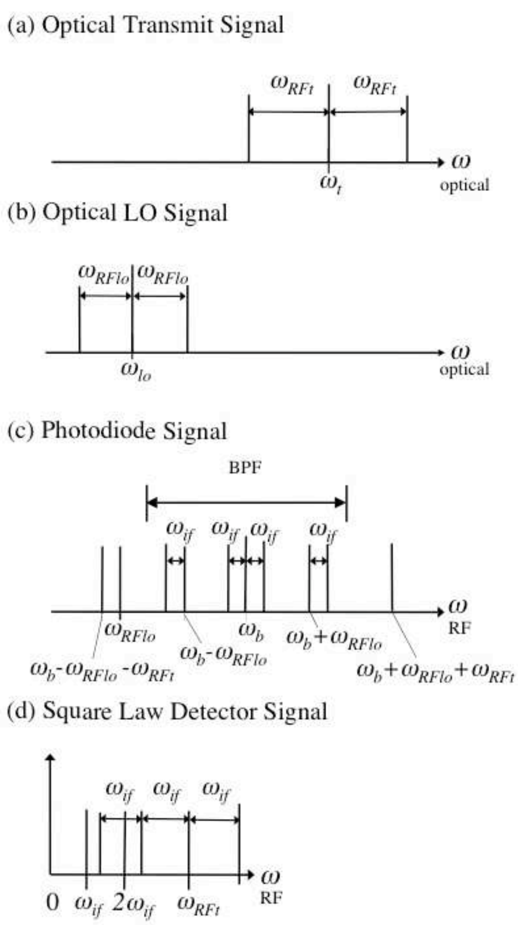

Fig. 2. Spectra of (a) the transmitted optical signal (side-bands of order $|n|>1$ not shown) (b) the modulated optical local oscillator (LO) signal (side-bands of order $|n|>1$ not shown) (c) the detected signal after the photodiode, and (d) the output signal from the square-law detector, including the relevant component at ω_{if}. Shown with DS code off. Not to scale.

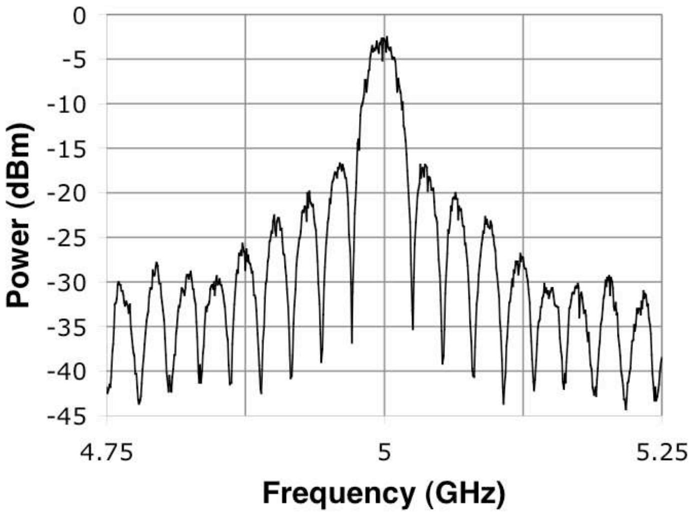

Fig. 3. Direct detection of transmitted signal in photodiode yields standard BPSK spectrum.

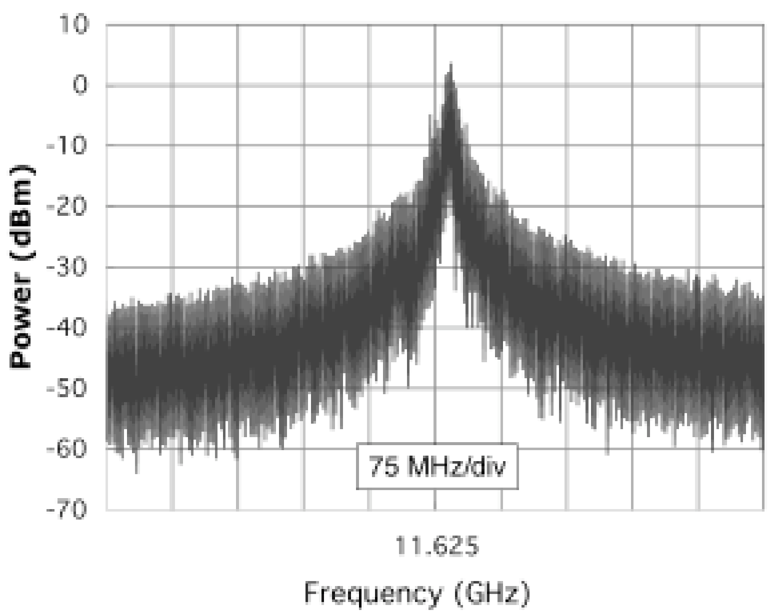

Fig. 4. Heterodyne signal at photodiode with DS signal off. Note the broad line-width.

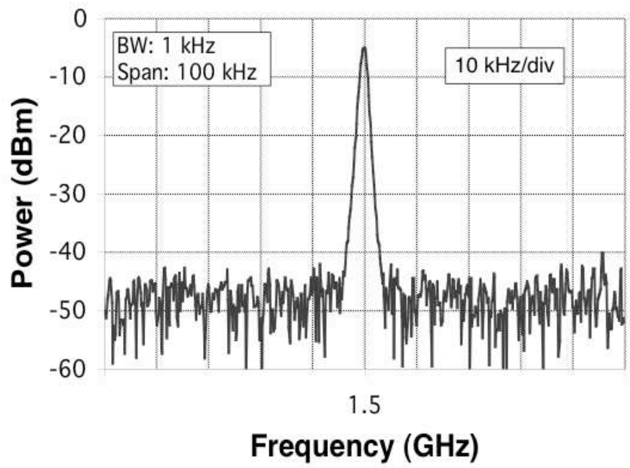

Fig. 5. Correlated signal following square law detector. DS code delay is set maximum signal. Note the narrow line-width.

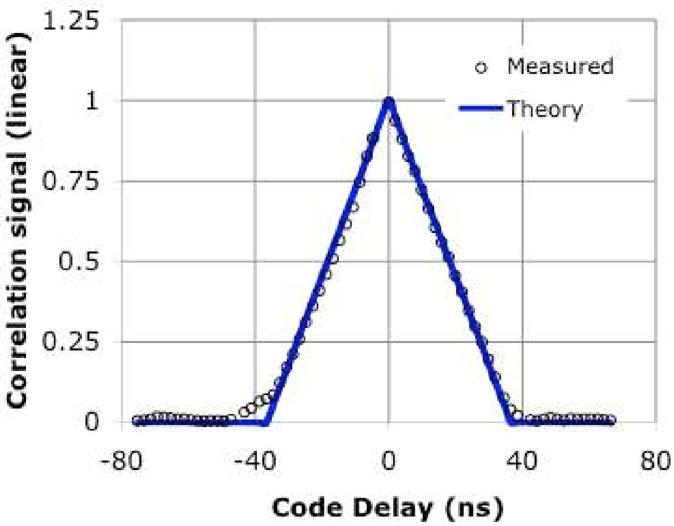

Fig. 6. IF signal amplitude (normalized) vs temporal offset of DS code. Measured signal and theoretical autocorrelation function of the *M*-sequence are both shown.