Wide-range, high-precision multiple microwave frequency measurement using a chip-based photonic Brillouin filter

Hengyun Jiang,1,2 David Marpaung,1,* Mattia Pagani,1 Khu Vu,3 Duk-Yong Choi,3 Steve J. Madden,3 Lianshan Yan,2 and Benjamin J. Eggleton1

1Centre for Ultrahigh Bandwidth Devices for Optical Systems (CUDOS), the Institute of Photonics and Optical Sciences (IPOS), School of Physics, University of Sydney, NSW 2006, Australia
2Center for Information Photonics & Communications, School of Information Science and Technology, Southwest Jiaotong University, Chengdu, Sichuan 610031, China
3Centre for Ultrahigh Bandwidth Devices for Optical Systems (CUDOS), Laser Physics Centre, Australian National University, ACT 2601, Australia
*Corresponding author: d.marpaung@physics.usyd.edu.au

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Spectrum analysis is a key functionality in modern radio frequency (RF) systems. In particular, fast and accurate estimation of multiple unknown RF signal frequencies over a wide measurement range is crucial in defense applications. Although photonic techniques benefit from an enhanced frequency estimation range along with reduced size and weight relative to their RF counterparts, they have been limited by a fundamental trade-off between measurement range and accuracy. Here, we circumvent this trade-off by harnessing the photon and phonon interactions in a photonic chip through stimulated Brillouin scattering, resulting in an accurate estimation of multiple RFs of up to 38 GHz with a record-low error of 1 MHz.

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1. INTRODUCTION

Fast and accurate estimation of unknown radio frequencies (RFs) is vital for antisteaeth defense, radar warning receivers, and electronic intelligence systems [1,2]. For these purposes, it is favorable to sense the frequency information of unknown intercepted signals before separately performing signal analysis and countermeasure [3]. However, modern electrical instantaneous frequency measurement (IFM) systems suffer from a limited measurement range of up to around 18 GHz; they are vulnerable to electromagnetic interference (EMI) and are incompatible with the concurrent processing of multiple frequencies [4]. Photonic measurement systems can overcome these limitations and extend the IFM receiver operation to a higher frequency range. Generally, photonic frequency measurement is realized by analyzing other easily measured parameters (e.g., RF or optical power) to estimate the frequency of an unknown RF signal, as shown in Fig. 1. Among these approaches, the power to frequency mapping is essential for the performance of an IFM system, such as the accuracy (i.e., the slope of mapping) and measurement range. Photonic IFM has shown potential for a wide measurement range of 50 GHz [5], a high accuracy of ±400 kHz [6], and can be integrated on-chip [7]. Additionally, the ability to measure simultaneous multiple frequencies is highly attractive in realistic spectrally cluttered environments [8]. However, these advantages have not been achieved simultaneously in a single system, due to the fundamental trade-off between measurement range and accuracy. Incorporating these features in one integrated on-chip system represents a significant step toward an advanced IFM receiver.

In this article, we demonstrate experimentally a frequency measurement system that simultaneously achieves the estimation of multiple frequency measurement up to 38 GHz with errors lower than 1 MHz in a centimeter-scale chalcogenide glass waveguide. We achieve this performance using distributed RF power-to-frequency mapping [see Fig. 1(b)], in which the RF power-to-frequency mappings are separately built in different frequency bands. Our approach circumvents the fundamental trade-off between measurement range and accuracy. Its channeledized frequency band offers an inherent capability to resolve and process multiple simultaneous RF inputs over a wide spectrum, provided the separation between multiple microwave frequencies is larger than the channel frequency spacing. The enabling technology for this breakthrough is the recently reported microwave photonic filter based on a very narrow linewidth and high extinction of stimulated Brillouin scattering (SBS) [9,10]. The results presented here point to new possibilities for creating a
high-performance integrated on-chip IFM system that will help assure critical mission success at minimal costs and enhanced security for manned and/or unmanned aircraft and surface vessels and next-generation radar, with potential for monolithic integration in silicon chips [11,12].

2. PRINCIPLE OF FREQUENCY MEASUREMENT

Figure 2(a) shows the conceptual diagram of our proposed IFM approach. The unknown RF in our proposed IFM approach is determined by two parameters: the coarse estimation of the frequency band and the higher precision. This allows the unambiguous determination of the unknown frequencies from the RF inputs, by means of the RF power change. The key to our approach is the SBS microwave photonic bandstop filter with a central frequency of \( f_c \), a narrow bandwidth of tens of MHz, and an anomalously high suppression (>60 dB) [9,10]. In this scheme, the SBS optical pump is generated by the dual-sideband suppressed-carrier modulation of the laser by the RF signal to be measured. The generated transfer function of the filter with single or multiple stop bands at a central frequency of \( f_{unknown} - \Omega_B \) (\( \Omega_B \) is the SBS frequency shift) is formed spontaneously. By subsequently launching a reference signal with a fixed frequency step \( (Nf_s, N = 1, 2, \ldots, n, n+1) \) at the input of the SBS filter, the output RF power of these frequency tones (channels) is changed by the stop band.

By specifically selecting the frequency step of the reference signal \( f_{ref} \) to be half of the bandwidth of the stop band, we ensure that there are only two adjacent channels [for example, \( nf_s \) and \( (n+1)f_s \)] that undergo a significant change in RF power and are located separately at both sides of the central frequency in the presence of one unknown frequency tone at most. Therefore, the central frequency \( (f_c = f_{unknown} - \Omega_B) \) of the SBS bandstop filter must be located between the two adjacent channels. This will give the coarse estimation of the unknown frequency, i.e., the frequency band of the unknown frequency is \( nf_s + \Omega_B < f_{unknown} < (n+1)f_s + \Omega_B \).

The RF power changes at the two channels offer more accurate information about the unknown frequency in our IFM approach. In order to understand the relation between the unknown RF and the RF power change in each channel, one can suppose that the central frequency of the filter can be gradually increased, the power change of a single channel (e.g., the channel of \( nf_s \)) is recorded accordingly [see the dashed line in Fig. 2(b)]. This RF power change response with respect to the central frequency of the filter (i.e., \( f_c \)) is the copy of the stop band transfer function with the central frequency at \( nf_s \). Hence, the analytical output RF power change responses at different frequency channels, i.e., \( \Delta P(Nf_s, f_c) \) \( (N = 1, 2, \ldots, n) \) are obtained as shown in Fig. 2(b). Upon an inspection of the frequency band of the unknown in Fig. 2(b), we found that the responses of the output RF power change at adjacent channels of \( nf_s, (n+1)f_s \) can separately offer a RF power to frequency mapping for higher frequency estimation. These two mappings are a pair of complementary functions and, thus, a linear amplitude comparison function (i.e., ACF) can be formed by \( \Delta P(nf_s, f_c) - \Delta P((n+1)f_s, f_c) \) [dB] in the frequency band of \( [nf_s, (n+1)f_s] \) as shown in Fig. 2(c). Such an ACF will provide a linear mapping between the central frequency of the filter (or the unknown frequency) and the power ratio of the adjacent channel. After comparing with the mapping in Fig. 2(b), it can offer a larger slope in the power-to-frequency mapping and ensure equal precision over the entire range of measurement.

Higher precision for estimating the unknown frequency can be achieved by comparing the ratio of the measured RF powers in adjacent channels (i.e., \( a-b \)) with the ACF of the system. Incidentally, the subtraction of the measured power values can cancel the additive noise along with the immunity to power fluctuations [7]. Therefore, the unknown frequency can be determined by \( nf_s + \Omega_B < f_{unknown} < (n+1)f_s + \Omega_B \) and \( f_{unknown} = ACF^{-1}(a-b) + \Omega_B \) in our approach.

Further, using only two adjacent channels for a frequency tone, we can simultaneously measure multiple frequency tones. The choice of the frequency step \( f_{ref} \) is important because the transfer function of the bandstop filter is symmetrical, in which each amplitude value has two frequency values associated with it. Hence, we have to measure the power change at two adjacent channels to unambiguously determine an unknown central frequency (i.e., \( f_c \) is chosen to be half of the bandwidth of the stop band). In our IFM approach, the measurement range is determined by the frequency range of the reference signal and the accuracy is defined by the transfer function of the bandstop filter. However, in order to monitor the output RF power at each channel of the reference signal, each channel has to be separated and then the RF power is measured using parallel RF power meters. This can make the system very complex and expensive. Fortunately, the technique of a frequency-scanning receiver offers an efficient solution to this problem. It sweeps a wide frequency bandwidth at a very high time speed, and then converts the frequency information into time data. Thus, only one power meter is required to measure the power change at different channels.
The sweep speed of the photonic scanning receiver is of the order of nanoseconds [13], while the maximum speed of sweep has to be slower than the phonon lifetime in the material (e.g., ~20 ns in our experiment), to ensure efficient build-up of SBS.

3. EXPERIMENT

A. Designed IFM System

Based on the approach above, we design a photonic IFM system by combining the frequency-scanning receiving technique as shown in Fig. 3. The output of a laser with power of 20 dBm at 1549.9 nm is split into two paths using an optical coupler. By feeding a scanning-frequency RF signal (≤600 μs switching speed in our experiment; Agilent N5183A) with a fixed frequency step (chosen to be half of the Brillouin linewidth, νB) to a dual-parallel Mach–Zehnder modulator (DPMZM) through a 90° hybrid coupler, the probe signal of SBS is generated on one path, creating a nearly phase-modulated signal where the optical sidebands are out of phase but have unequal amplitudes [Fig. 3(b)]. On another path, a microwave signal with the unknown frequency (funknown) is modulated into a carrier-suppressed double-sideband signal using a Mach–Zehnder modulator (MZM) as Fig. 3(c). The output signals of the optical modulators are properly amplified using two Er-doped fiber amplifiers working at the automatic power control mode. Hence, unknown RF signals with different RF powers will induce SBS pumps with uniform optical powers.

The amplified double-sideband signal is launched into a 6.5-cm-long As2S3 rib optical waveguide via a lens-tipped fiber as the pump of the SBS process, and the amplified probe signal is injected into the waveguide from the opposite end. The waveguide used had a cross section of (0.85 μm × 4 μm), a mode area of 2.3 μm², and a high SBS gain coefficient (g₀ ~ 0.74 × 10⁻⁹ m/W) [14]. The total inserted loss of the waveguide is about 12 dB, including the coupling loss and a propagation loss of 0.3 dB/cm. In this nonlinear waveguide, the weaker sideband of the probe signal can be amplified by the SBS gain spectrum and the stronger sideband can be reduced by the SBS loss spectrum [Fig. 3(d)]. We assure that the two sidebands would be totally balanced by the SBS gain and loss spectra only at the center of the SBS resonance. Upon photodetection, the beat signals generated from the optical carrier and the two sidebands cancel perfectly, whereas the signal outside the SBS resonance does not cancel completely due to the amplitude difference between the sidebands. This forms a desired bandstop filter with a central frequency of funknown − ωνp, a narrow bandwidth of νB, and an anomalously high suppression in our scheme [Fig. 3(e)]. Thus, the moving reference signal is filtered by the stop band with the unknown central frequency, showing different RF power changes at different channels at the output of the photodetector.

To experimentally build this IFM system, we measure the carrier-suppressed double-sideband signal at the output of the MZM as shown in Fig. 4(a), which acts as two pumps of the SBS process. By properly adjusting the three bias voltages of the DPMZM via a programmable multichannel voltage supply with 1 mV voltage accuracy (Hameg HM7044G), the optical spectra of the nearly phase-modulated signal with unequal amplitude is measured at a test frequency (~20 GHz) as shown in Fig. 4(b). We formed a SBS filter with a bandwidth (νB) of ~50 MHz and a high stop band rejection of ~36 dB. The SBS frequency shift of the waveguide is measured to be ΩB ≈ 7.695 GHz. Subsequently, the RF power-to-frequency mapping is established before starting the estimation of the unknown frequency. Due to the complex transfer function of the SBS filter, the ACF is established by precisely measuring the continuous RF power distribution as a function of RFs for the adjacent channels. The ratio of these two continuous functions in a spectral band yields a linear ACF. The frequency of the reference signal is set to be ~2,305 GHz; the function of the output RF power related to the modulated frequency of the MZM is measured by continuously adjusting the modulated frequency of the MZM as shown in Fig. 5(a) (blue line). It resembles a copy of the transfer function of the SBS filter with the central frequency at ~10 GHz (i.e., 2,305 GHz + ΩB). Here, the frequency step of the reference signal is set to be ~25 MHz (i.e., half of the bandwidth). Figure 5(a) shows the function of the output RF power at frequencies of ~2,330 and ~2,355 GHz of the reference signal. Thus, the relationship between the modulated frequency of the MZM (or the unknown frequency) and the output RF power, NνB/2 (N = 1, 2, 3, …), is demonstrated in our system.

Fig. 3. Experimental setup. LD, laser; OC, optical coupler; MZM, Mach–Zehnder modulator; DPMZM, dual-parallel MZM; EDFA, Er-doped fiber amplifier; ISO, inline optical isolator; PD, photodetector. (a)–(e) The optical/RF spectra at different points.

Fig. 4. Measured optical spectra. (a) Carrier-suppressed double sideband at the output of the MZM and (b) phase-modulated signal with unequal amplitude at the output of the DPMZM.
and 13.046 GHz; the power change at four channels are obtained at 10.005 and 10.090 GHz, 10.097 and 10.168 GHz, and 10.013 GHz. The power ratio at these two channels is about 21.07 dB. We observed that the mapping between the modulated frequency of the MZM and the output RF power ratio is approximated as a linear function in each frequency band. The ACF has a very high slope of ~70 dB over a 25 MHz frequency band. The distributed RF power ratio-to-frequency mapping is revealed and an averaged resolution of frequency of ~2.8 dB/MHz is achieved in our chip-based system.

B. Unknown Frequency Estimation

The first group in Fig. 6 shows the RF powers measured at different channels when the unknown frequency is set as 10.002 GHz. As expected, the channels detuned from the stop band were hardly affected, whereas the output power at frequencies of 2.305 and 2.330 GHz of the reference signal (in the stop band) is changed. Subsequently, the frequency of the unknown signal is coarsely estimated into the frequency band of 10.000–10.025 GHz. The power ratio at these two channels is about ~21.07 dB, and the measured precise frequency of these inputs is ~10.0016 GHz after comparing with the ACF in Fig. 5. To verify the capabilities of this technique to measure multiple frequencies, we launched two microwave signals with unknown frequencies. The unknown frequencies are separately set as 10.005 and 10.090 GHz, 10.097 and 10.168 GHz, and 10.013 and 13.046 GHz; the power change at four channels are obtained and every two channels of power change are adjacent for each case, and the measured precise frequencies of these inputs are found to be ~10.0048 and 10.0897 GHz, 10.0969 and 10.1682 GHz, and 10.0136 and 13.0456 GHz by using the same method with a single frequency tone. The minimum frequency interval of multiple frequency tones could be ~50 MHz in our experiment, so that the SBS resonances do not overlap. The minimum measured frequency of our system should be higher than the SBS frequency shift of the material. The approach for unknown frequency measurement is demonstrated successfully and the ability to independently estimate multiple frequency tones is verified in our experiment.

Moreover, the estimation error of frequency is measured to evaluate the performance of our chip-based IFM system using distributed RF power ratio-to-frequency mapping. Figure 7 shows that the estimation error is of the order of ±1 MHz over a wide measurement range of 9–38 GHz in our system. The measured frequency estimation of an RF signal is located at 13 frequency bands.

4. DISCUSSION

We compare the important performance of this chip-based IFM system to those of other state-of-the-art IFM techniques in Table 1. The SBS-on-chip IFM system uniquely combines the advantages of two different IFM systems: (1) the wide measurement range (20 GHz) and multiple RF input frequency processing of the scanning receiving technique and (2) the high accuracy (±400 kHz) of the ACF technique based on narrow-band filters. The ratio of measurement range to accuracy is applied to comprehensively evaluate the performance of different IFM systems. This system achieves a 75-fold improvement when compared with the best-performance photonic IFM system reported previously. Moreover, the performance can be improved by reducing the insertion loss of the chip, through the use of on-chip tapers.

Table 1. Performance Comparison of Existing IFM Systems

<table>
<thead>
<tr>
<th>Technology</th>
<th>Range (GHz)</th>
<th>Accuracy (MHz)</th>
<th>Input</th>
</tr>
</thead>
<tbody>
<tr>
<td>SBS on chip (this work)</td>
<td>9–38</td>
<td>1</td>
<td>M</td>
</tr>
<tr>
<td>Si3N4 filter on chip [7]</td>
<td>0.5–4</td>
<td>936</td>
<td>S</td>
</tr>
<tr>
<td>Delay loop [6]</td>
<td>6.94–6958</td>
<td>04</td>
<td>S</td>
</tr>
<tr>
<td>Scanning receiver [8]</td>
<td>0.1–20</td>
<td>250</td>
<td>M</td>
</tr>
<tr>
<td>Compressive sampling [5]</td>
<td>1–50</td>
<td>600</td>
<td>M</td>
</tr>
<tr>
<td>Digital IFM (electronic) [4]</td>
<td>0.5–18</td>
<td>2.4 (RMS)</td>
<td>S</td>
</tr>
<tr>
<td>Delay line (electronic) [16]</td>
<td>2–4</td>
<td>1</td>
<td>S</td>
</tr>
</tbody>
</table>

M, multiple; S, single.
for example. Our recent demonstration of fiber-based IFM indicates that superior performance can be achieved [15]. In this experiment, improved performance was achieved because of a lower insertion loss, a reduced SBS linewidth, and improved stability. When the insertion loss of the system is reduced, a stop band rejection of >60 dB can be achieved in the SBS filter. Thus, we can obtain a steeper power-to-frequency mapping for more accurate frequency measurement. The performance of our system is theoretically defined by the formed distributed power-to-frequency mapping using the SBS filter, which is limited by the SBS waveguide we used. At present, the main limitation to higher performance is the stability of the fiber-to-chip coupling in our experiment.

In contrast to traditional RF-scanning receivers or electrical spectrum analyzers (ESAs), our technique reveals a distinct advantage. In ESAs, the accuracy is dictated by the width of the scanned bandpass filter. This, in turn, poses a trade-off between accuracy and acquisition time for wide frequency range measurements. In our technique, the resolution is determined by the slope of the bandstop filter. By using a very high slope, we have achieved very high resolution (~1 MHz) with much large frequency steps (~25 MHz). This greatly relaxes the tight constraints of accuracy and acquisition time. The time to step the frequency can be as low as tens of nanoseconds in our technique.

5. CONCLUSION

In conclusion, we have measured multiple gigahertz (GHz) microwave frequencies of up to 38 GHz with a low error of <1 MHz in a centimeter-scale chalcogenide glass waveguide. The unique mapping of distributed RF power ratio-to-frequency mapping, which successfully breaks through the trade-off between accuracy and measurement range, is formed in a very simple system. Our approach combines the frequency agility of a scanning receiver and the accuracy of ACF based on the photon and phonon interactions through SBS. Our demonstration opens the path toward monolithic integrated IFM with a tens of GHz measurement range, sub-megahertz accuracy, and multiple frequency measurement capability in silicon chips.

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