DOI: 10.35840/2631-5092/4538

International Journal of
Optics and Photonic Engineering

ISSN: 2631-5092

# A Scheme of Instantaneous Frequency Measurement with High Precision Assisted by Photonics

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**Abstract**

In modern complex battlefield environment, it is important to obtain the frequency spectrum precisely. We propose an operational method of instantaneous frequency measurement (IFM) assisted by photonics, which can achieve high precision based on stimulated Brillouin scattering (SBS), giving the credit to the narrow linewidth of gain spectrum of SBS. We use MZM and DPMZM cascade to generate a tunable continuous optical signal and modulate the measured signal to the continuous optical signal and a DPMZM is used to generate pump light, the two beams of light are injected into the fiber, the stimulated Brillouin scattering (SBS) effect occurs in the optical fiber and the Brillouin gain spectrum (BGS) is formed. By setting the scanning frequency, the Brillouin gain varies with frequency and the amplitude comparison function (ACF) can be formed in the narrow line band of BGS. And the measurement of full frequency band is realized through a reference signal, the measured frequency range is limited only by photoelectric device. Estimation of multiple radio-frequency (RF) signals can also be achieved with a resolution of 250 MHz. In the numerical simulation, the average measurement error less than 1 MHz is achieved in this scheme.

**Index Terms**

Instantaneous frequency measurement, Photonics, High precision, Stimulated Brillouin scattering

## Introduction

In modern electronic warfare (EW) applications, frequency measurement has played an important role. Its measurement range is limited by the bandwidth of electronic components, and it is more and more difficult for electronic methods to achieve high frequency measurement precisely with the increasing frequency [1]. In recent years, with the development of microwave photonics technology, researchers has proposed to achieve frequency measurements assisted by microwave photonics [2-5] and there are mainly three types of frequency measurement based on photon, which are based on time-domain scanning [6,7], optical-frequency-comb-assisted channelized receivers [8-12] and frequency-to-power mapping [4,13-19]. The method of frequency-to-power mapping, which is to map the frequency of the microwave signal to microwave power [20,21] or optical power [13,22], attracts a large number of researchers.

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**Accepted:** August 16, 2021; **Published:** August 18, 2021



Li et al. Int J Opt Photonic Eng 2021, 6:038

**Citation:** Li J, Xu P, Song K, Du K, Lu M, et al. (2021) A Scheme of Instantaneous Frequency Measurement with High Precision Assisted by Photonics. Int J Opt Photonic Eng 6:038

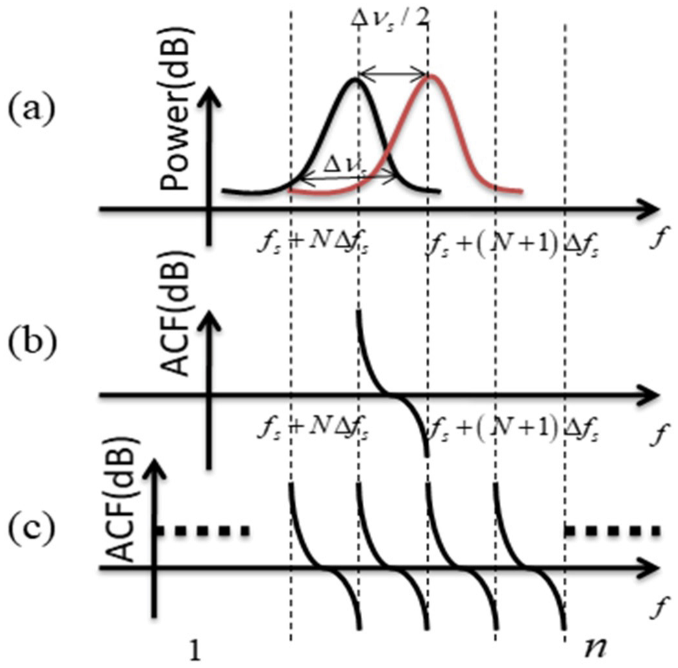


**Figure 1:** The schematic diagram of the proposed scheme.

The relationship between the frequency and the power established by any principle such as filter or nonlinear effect is monotonic function, so the frequency can be obtained by measuring the power. Each of the three approaches has its own advantages. The method of frequency-to-power mapping achieved a wide measurement range and a high accuracy with a simple architecture [4,13-19]. Especially, combined with the narrow linewidth of gain spectrum of the stimulated Brillouin scattering effect, the accuracy can be further improved theoretically [23]. Fiber dispersion or microwave filter response can all construct a monotone function flexibly [10,24,25]. The performance of an IFM system, such as the accuracy and measurement range, has achieved a wide measurement range and a high accuracy [26,27], and can be integrated on-chip [23]. However, these advantages have not been achieved simultaneously in a simple system. In the integrated system, the scanning range is limited by the bandwidth of the modulator, there will be the disadvantage that the frequency measurement range cannot be fully covered within the afforded range of the scanning range. At present, the error of most frequency measurement schemes assisted by photonics is around the hundreds of MHz and the error of using SBS can be around 1 MHz and the frequency measurement range is always limited by the scanning range in SBS scheme.

In this article, we numerically investigate the operational principle of full frequency measurement based on stimulated Brillouin scattering. The idea can achieve a measurement error lower than 1 MHz within the bandwidth of photoelectric device and when a large bandwidth signal is measured, only a small scanning range signal is required. We achieve this performance using distributed RF power-to-frequency mapping based on SBS shown in Figure 1, in which the RF power-to-frequency mappings are separately built in different frequency bands with a large slope. As we all known, a steep slope means high precision. Meanwhile, a reference signal, whose frequency is larger than Brillouin frequency shift (BFS), is used to achieve the full frequency measurement, which adopt to different BFSs in other system.

## Principle

The schematic diagram of our proposed scheme is shown in Figure 1. The unknown frequency band can be coarsely estimated by means of the RF power change of different bands. Accurate measurement in any bands requires the use of SBS microwave photonics bandpass filter with a narrow bandwidth of tens of MHz and an anomalously high suppression. The generated transfer function of the filter with single or multiple pass bands at a central frequency of $f_c + f_r + f_x - \nu_B$ ($\nu_B$ is the local BFS) shown as Figure 2. Due to the reference

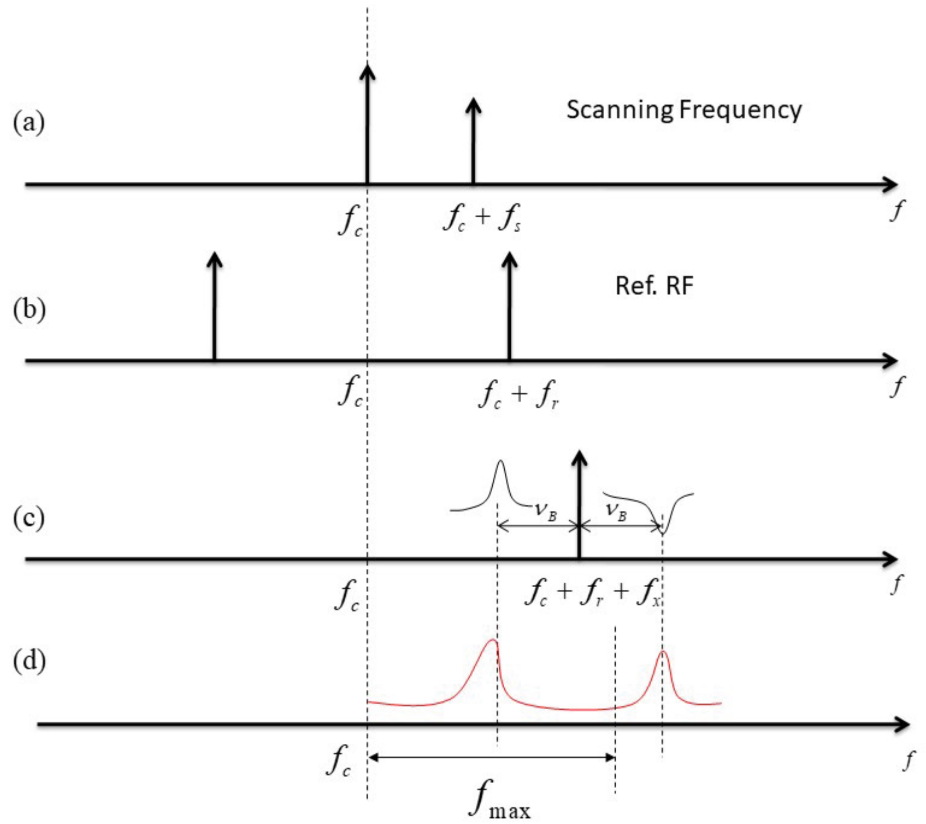


**Figure 2:** Diagram of SBS filter formation.

signal $f_r$, the frequency measurement range is fully covered and the frequency range is no longer limited, which is different from others. By launching a scanning signal ($f_s$) with a fixed frequency step ($\Delta f_s$) at the input of the SBS filter, the output RF power of these frequency channels is changed by the stop band. In order to ensure that there are only two adjacent channels without spectrum loss, the frequency step of $\Delta f_s$ should be half of the bandwidth of the zero point ($\Delta \nu_S$) shown as Figure 1a. The coarse estimation of the unknown frequency will be located in the frequency band of $f_s + N\Delta f_s - f_r + \nu_B \le f_x \le f_s + (N+1)\Delta f_s - f_r + \nu_B$ ($N = 1, 2, 3 \ldots n$). From Figure 1a, we found that the responses of the output RF power change at adjacent channels of $f_s + N\Delta f_s$, $f_s + (N+1)\Delta f_s$ can separately offer a RF power to frequency mapping with a pair of complementary functions. Thus, a linear amplitude comparison function (ACF) can be formed by $\Delta P(f_S + N\Delta f_S) - \Delta P(f_S + (N+1)\Delta f_S)$ in the frequency band of [$f_s + N\Delta f_s, f_s + (N+1)\Delta f_s$] as shown in Figure 1c. Such an ACF will provide a liner relationship between the unknown frequency ($f_x$) and the power ratio of the adjacent channel. The subtraction of the measured power values can cancel the additive noise along with the immunity to power fluctuations [23]. In order to achieve full coverage of the frequency range, the frequency of reference signal should be higher than BFS to make the gain spectrum away from the carrier. Assuming that the photodetector bandwidth is limited to $f_{\max}$ shown as Figure 2, the frequency of reference signal should also be higher than $f_{\max} - \nu_B$ to eliminate the interference of loss SBS spectrum. Therefore, the unknown frequency can be determined by $f_x = ACF^{-1}(a-b) - f_r + \nu_B$ in our approach, where $a$ and $b$ represent RF powers in adjacent channels respectively. A frequency-scanning receiver sweeps a wide frequency bandwidth at a very high time speed, and to ensure efficient build-up of SBS, the maximum speed of frequency-scanning receiver has to be slower than the phonon lifetime in the material (~20 ns) [23].

## Simulation and Discussion

The scheme of frequency measurement we proposed is shown as Figure 3. The output light of a laser is split into two paths through an optical coupler (OC). A scanning-frequency RF signal ($f_s$)

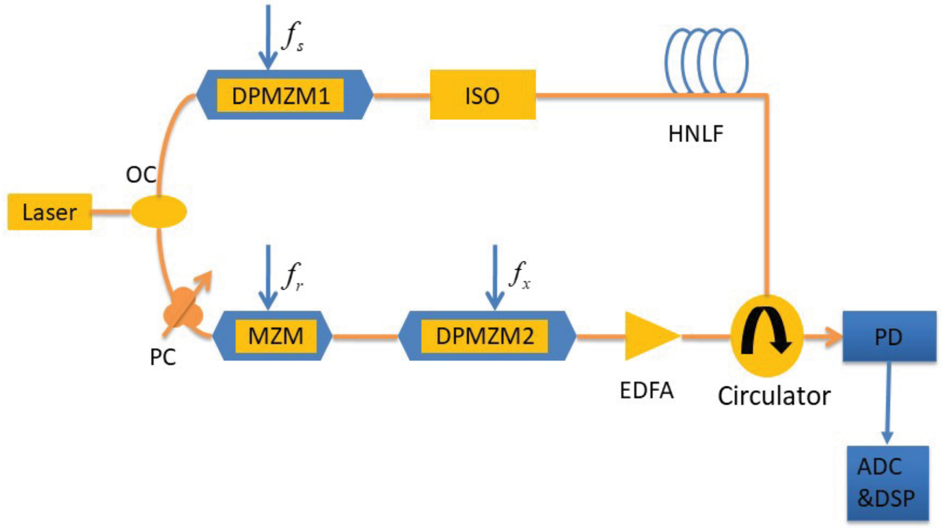


**Figure 3:** The system diagram.

OC: Optical Coupler; PC: Polarization Controller; ISO: Isolator; HNLF: High Nonlinear Optical Fiber; MZM: Mach-Zehnder Modulator; DPMZM: Double Parallel Mach-Zehnder Modulator; EDFA: Erbium-Doped Optical Fiber Amplifier.

with a fixed frequency step $\Delta f_s = \Delta \nu_S / 2$ based on the approach above is fed into a dual parallel Mach-Zehnder modulator (DPMZM1) through a 90° hybrid coupler achieving carrier suppression SSB modulation. On another path, a reference microwave signal $f_r$ is modulated into a carrier-suppressed double sideband signal using a Mach-Zehnder modulator (MZM) as Figure 2b. The unknown frequency $f_x$ is fed into a dual parallel Mach-Zehnder modulator (DPMZM2) cascaded with MZM to achieve carrier suppression SSB modulation assisted by a 90° hybrid coupler as Figure 2c. The output signal of the DPMZM2 is properly amplified using an Er-doped fiber amplifier. Hence, unknown frequency signals ($f_x$) will induce SBS effect with power change in the high nonlinear optical fiber (HNLF). This forms a desired SBS passband filter with a central frequency of $f_c + f_r + f_x - \nu_B$ and a narrow zero-point linewidth of $\Delta \nu_S$. Based on the approach above, the unknown frequency signal can be expressed as

$$f_x = f_s - f_r + \nu_B \tag{1}$$

Where the instant scanning frequency ($f_s$) can be obtained by ACF. In the scheme proposed in this paper, ACF curve can be fitted by means of three-order Hermite interpolation.

In this paper, the upper sideband of the scanning signal is selected to be processed by SBS. The Brillouin gain and loss can be expressed as [19]:

$$g(f) = \frac{g_0}{2}\frac{(\Delta\nu_B/2)^2}{f^2+(\Delta\nu_B/2)^2} + j\frac{g_0}{4}\frac{\Delta\nu_B f}{f^2+(\Delta\nu_B/2)^2} \tag{2}$$

$$a(f) = -\frac{g_0}{2}\frac{(\Delta\nu_B/2)^2}{f^2+(\Delta\nu_B/2)^2} - j\frac{g_0}{4}\frac{\Delta\nu_B f}{f^2+(\Delta\nu_B/2)^2} \tag{3}$$

Where $\Delta\nu_B$ represent the SBS 3 dB-linewidth, $g_0 = g_B I_p L_{eff} / A_{eff}$, $g_B$ is the line center gain, $I_p$ is the power of pump wave, $L_{eff}$ and $A_{eff}$ are effective fiber length and effective mode area of DSF respectively.

Considering SBS effect, the optical signal after circulator can be expressed as

$$E(t) \propto \begin{cases} J_0(m)\exp(2\pi f_c t) \\ +J_1(m)\exp\begin{cases} \left\{ g\left[\left(f_p - \nu_B\right) - \left(f_c + f_s\right)\right] - a\left[\left(f_p + \nu_B\right) - \left(f_c + f_s\right)\right]\right\} \\ +j(2\pi f_s t) \end{cases} \end{cases} \tag{4}$$

Where $f_p$ is equal to $f_c + f_r + f_x - \nu_B$ based on the approach above, the Eq.4 can be expressed as

$$E(t) \propto \begin{cases} J_0(m)\exp(2\pi f_c t) \\ +J_1(m)\exp\begin{cases}\{g[f_r+f_x-\nu_B-f_s]-a[f_r+f_x+\nu_B-f_s]\} \\ +j(2\pi f_s t)\end{cases}\end{cases} \tag{5}$$

Omitting the dc and the other harmonic components, the optical power input into the PD is expressed approximately as:

$$P \approx 2J_0(m)J_1(m)\left\{G(f_s)A(f_s)\cos\left[2\pi f_s t+\phi_g(f_s)+\phi_a(f_s)\right]-\cos 2\pi f_s t\right\} \tag{6}$$

According to Eq.(2) and Eq.(3), it can be deduced that

$$\begin{aligned} G(f_s) &= \exp\left\{\mathrm{Re}\left[g(f_r+f_x-\nu_B-f)\right]\right\} \\ &= \exp\left\{\frac{g_0}{2}\frac{(\Delta\nu_B/2)^2}{(f_r+f_x-\nu_B-f)^2+(\Delta\nu_B/2)^2}\right\} \end{aligned} \tag{7}$$

$$\begin{aligned} A(f_s) &= \exp\left\{\mathrm{Re}\left[a(f_r+f_x+\ \ B-f_s)\right]\right\} \\ &= \exp\left\{-\frac{(\qquad /2)}{(f_r+f_x+\nu_B-f_s)^2+(\Delta\nu_B/2)^2}\right\} \end{aligned} \tag{8}$$

$$\begin{aligned} \phi_g(f_s) &= \mathrm{Im}\left[g(f_r+f_x-\nu_B-f)\right] \\ &= \frac{g_0}{4}\frac{\Delta\nu_B(f_r+f_x-\nu_B-f)}{(f_r+f_x-\nu_B-f)^2+(\Delta\nu_B/2)^2} \end{aligned} \tag{9}$$

$$\begin{aligned} \phi_a(f_s) &= \mathrm{Im}\left[a(f_r+f_x+\nu_B-f_s)\right] \\ &= -\frac{g_0}{4}\frac{\Delta\nu_B(f_r+f_x+\nu_B-f_s)}{(f_r+f_x+\nu_B-f_s)^2+(\Delta\nu_B/2)^2} \end{aligned} \tag{10}$$

The output electric field after detecting can be expressed as:

$$\begin{aligned} E_{out}(t) &= \Re\langle P\rangle \\ &\propto G(f_s)A(f_s)\cos\left[2\pi f_s t+\phi_g(f_s)+\phi_a(fs)\right]-\cos(2\pi f_s t) \end{aligned} \tag{11}$$

Where $\Re$ means the responsivity of PD to the input optical power.

In this paper, the reference signal is tunable, which can eliminate the interference of loss SBS spectrum associating with the bandwidth of the photodetector shown as Figure 2c ~ Figure 2d. In the numerical simulation, assuming that $g_0 = 5$, $\nu_B = 9.2$ GHz, $\Delta\nu_B = 40$ MHz, which induces that the $\Delta\nu_S = 500$ MHz. So that we set the scanning frequency step is $\Delta f_s = 250$ MHz based the approach above. When the $f_s$ scans from 0-20 GHz (where the bandwidth of the PD is $f_{max} = 20$ GHz.) with the step of 250 MHz, the several formed channels of ACF is shown as Figure 4 ~ Figure 5 with $f_r = 20$ GHz. In each channel, the steeper the slope of ACF curve is, the higher the accuracy of frequency measurement is. As can be seen from the Figure 4 to Figure 5, the SLOPE of the ACF curve is roughly the same for each channel. The ACF is calculated in each frequency band by three-order Hermite interpolation according to the measured functions of the output RF power. So power to frequency conversion can be achieved through ACF and the frequency of unknown signal can be obtained based on Eq.(1). Based on the approach above, Figure 6 indicates that the average error of measurement is less than 1 MHz using three-order Hermite interpolation on ACF. If the bandwidth of the photoelectric device is 40 GHz, the Figure 7 shows the channel from 20.000 GHz ~ 20.250 GHz when $f_r = 20$ GHz.

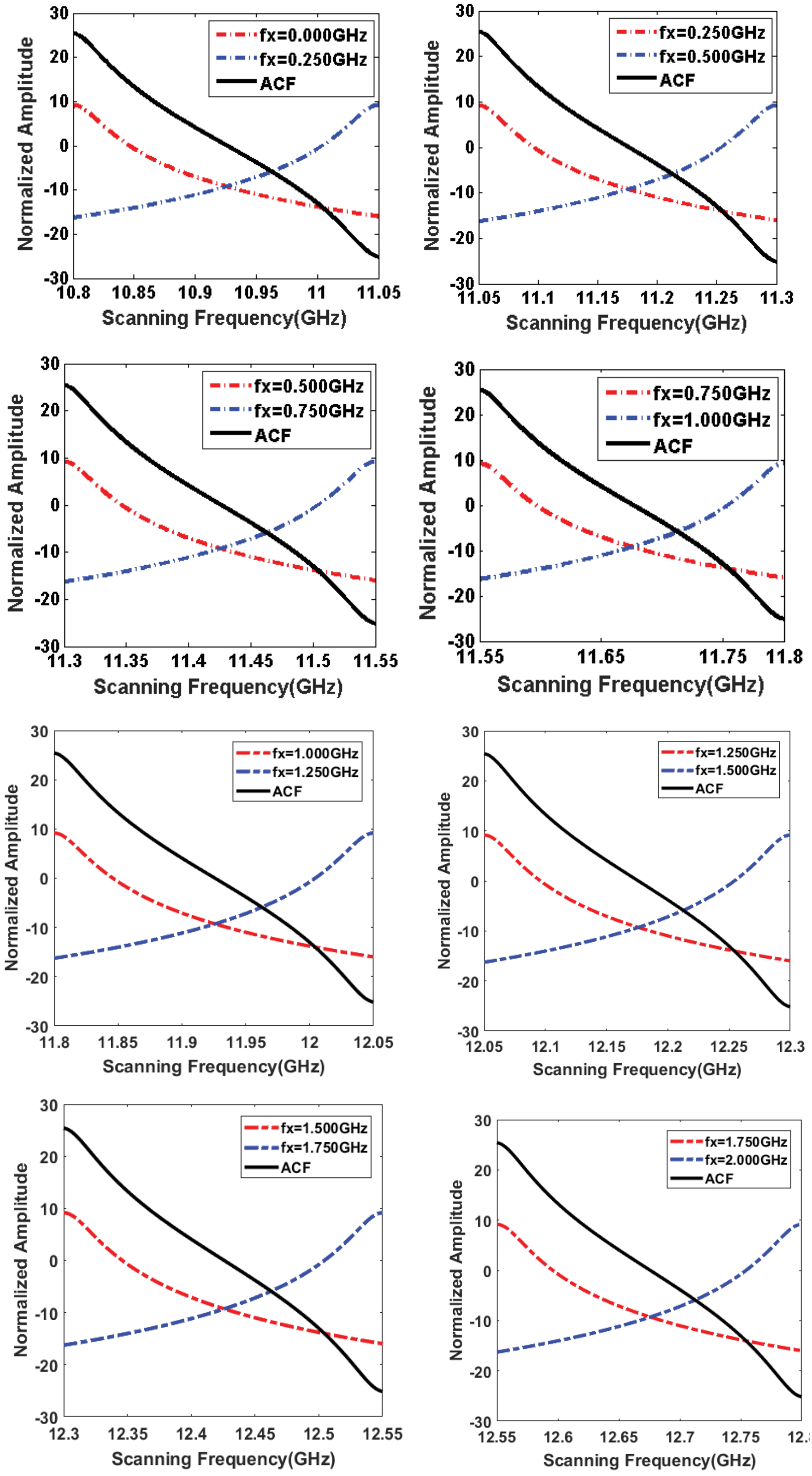


**Figure 4:** Channels from 0 ~ 2G.

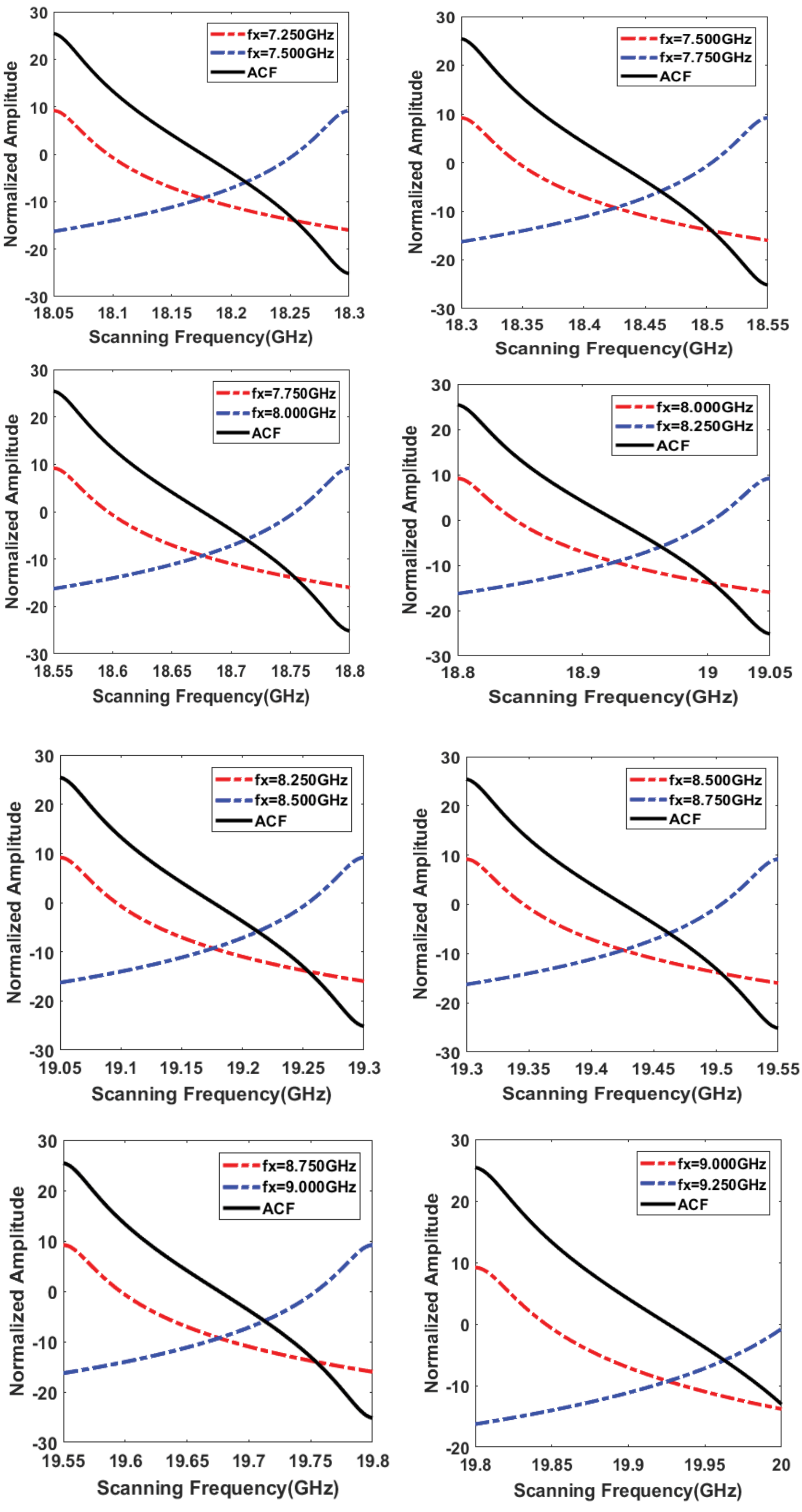


**Figure 5:** Channels from 7.25 GHz ~ 9.250 GHz.

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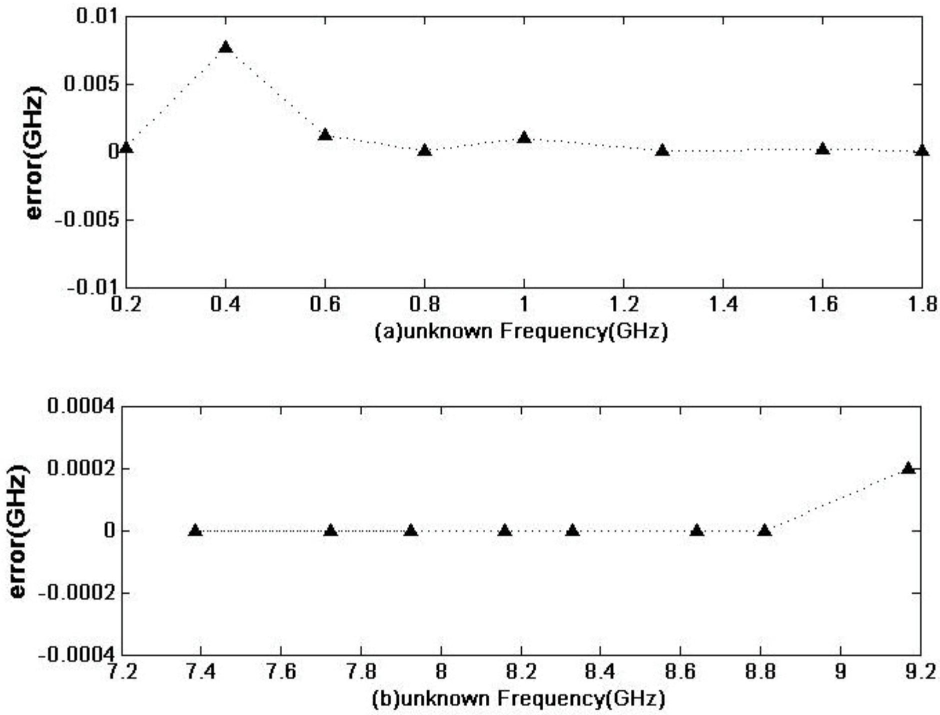


**Figure 6:** a) Measured frequency estimated errors of an RF signal in the range of 0-2.5 GHz; b) Measured frequency estimated errors of an RF signal in the range of 7.25 GHz-9.20 GHz

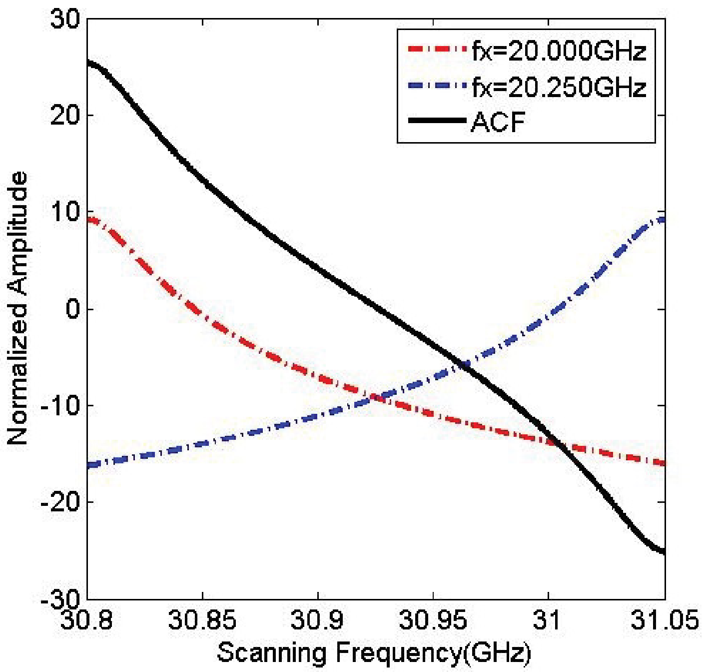


**Figure 7:** The channel from 20.000 GHz ~ 20.250 GHz when $f_r = 20$ GHz and the bandwidth of the photoelectric device is 40 GHz.

## Conclusion

In this paper, an instantaneous frequency measuring method with high precision assited by SBS is proposed, which can achieve the measurement of full frequency band without the interference of loss SBS through a tunable reference signal, and the measured frequency range is limited only by photoelectric device. Assuming that the bandwidth of photoelectric device is 40 GHz and the unknown signal may be in the range of 20.000 GHz-20.250 GHz, the ACF can be formed by adjusting the reference signal as 20 GHz shown as Figure 7. And an accurate estimation of multiple RFs can also be achieved with a resolution of 250 MHz by measuring power variation per channel [23]. If the waveguide material can be used to replace the high nonlinear fiber, the system integration will be greatly improved.

## Disclosure

The authors declare no conflicts of interest.

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DOI: 10.35840/2631-5092/4538

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