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**BROADBAND SUPERHETERODYNE TRACKING
CIRCUITS FOR MILLIMETER-WAVE
MEASUREMENTS**

BY

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**PLASMA PHYSICS
LABORATORY**

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BROADBAND SUPERHETERODYNE TRACING CIRCUITS
FOR MILLIMETER-WAVE MEASUREMENTS

by J. L. Doane

Abstract

Superheterodyne stimulus-response measurements can be made with high dynamic range, but are often sensitive to oscillator frequency drift and noise. The usual techniques for reducing this sensitivity often become impractical at millimeter-wave frequencies and above. This paper describes a feed-forward tracing technique and its application to single frequency millimeter-wave interferometry and doppler-shift scattering measurements on tokamak plasmas. Swept frequency transmission measurements can also be made with high dynamic range using this technique.

INTRODUCTION

This paper considers superneterodyne stimulus-response measurements, for which the transmitter and local oscillator are physically close together. The technique described herein is not suitable for instruments such as radiometers, for which the local source is not a discrete oscillator.

The simplest configuration for a stimulus-response measurement is the homodyne arrangement, which requires only one high frequency oscillator. A sample of the transmitted signal is used as a coherent reference for the receiver mixer. To reduce the effects of oscillator AM noise and $1/f$ noise in the mixer, some kind of frequency offset is usually introduced. Sawtooth phase modulation (serrodyne) and single sideband amplitude modulation (SSB) are commonly used to produce this offset.¹ Alternatively, direct sawtooth or pulse frequency modulation of the oscillator can produce large offsets, when the reference and signal paths differ considerably in length.^{2,3}

Homodyne configurations with only one high frequency oscillator have certain disadvantages. The modulators generally have narrow BW bandwidths, especially at millimeter wave lengths where the components are large compared to a wavelength. Direct frequency modulation of the oscillator introduces some noise and spurious AM modulations that limits the dynamic range. Simple homodyne arrangements, as well as those with double sideband modulators (DSB), require adjustment of the reference phase at each measurement frequency. Such arrangements are also incapable of determining

the sign of the doppler frequency shift in scattering measurements. This last drawback can be circumvented by use of an image reject receiver, which however also has a narrow RF bandwidth.

Superheterodyne measurements usually require extreme frequency stabilization^{4,5} or feedback synchronization of the two high frequency oscillators.⁶ This is done either with automatic frequency control (AFC) or with a phase lock loop. Without synchronization, large relative drift of the two oscillators would necessitate a large IF signal bandwidth, which in turn decreases the measurement signal to noise ratio. Relative frequency drift of the millimeter wave oscillators also increases the complexity of the signal processing required on the IF signal. With or without synchronization, measurements with two millimeter wave oscillators are ordinarily narrowband and require retuning at each new frequency. Swept frequency measurements are ordinarily impractical, because the free running frequencies of the two oscillators do not track closely enough for the difference frequency to remain within the hold-in range of the feedback loop.

The feed-forward tracing circuit described in this paper was originally designed to make high dynamic range interferometry measurements at 140 GHz without feedback synchronization. A multichannel doppler-shift scattering system was designed later for the same frequency using basically the same approach. In addition, the approach

described herein can be used to make swept transmission measurements with high dynamic range. A tracking generator such as is used for this purpose at lower microwave frequencies is impractical at millimeter wavelengths. Such generators presuppose the availability of a fixed high frequency oscillator, an ultra wideband RF mixer, and a swept oscillator operating at frequencies much higher than the measurement frequencies.⁷

Following a short description of the tracking circuit is a discussion of its application to present and possibly future systems, and then a consideration of how FM noise on the oscillators affects the minimum obtainable system noise.

DESCRIPTION

The main features of the tracing circuit can be easily described with reference to Fig. 1. For concreteness we will assume that the RF oscillators operate at millimeter wavelengths and that the system to be measured is a plasma. Due to frequency noise and drift, lack of resettability, difference in tracking, etc., the transmitter and local oscillator (LO) are offset from their nominal frequencies by ϵ and δ , respectively. The nominal intermediate frequency (IF) is f_i . The signal frequency may experience a doppler shift f_p as it passes through the plasma.

The components above the dashed line in Fig. 1, the millimeter wave (RF) components, are the same as would be needed for any oscillator synchronization scheme. The

output of auxiliary mixer M1 would be fed to a discriminator in an AFC system ("signal following superheterodyne") or to a phase lock synchronizer. The isolators I1 and I2 suppress the reflections from M1 and signal mixer M2. R1 through R3 are level set attenuators. I2 is mainly needed to prevent multiple reflections from affecting measurement accuracy, while I1 reduces the amount of transmitter power coupled into the path containing R2.

To avoid interference, the amount of transmitter power that reaches M2 without going through the plasma must be significantly less than the signal power coming from the plasma (when $f_p = 0$). The losses in R1 through R3 and the directivities of couplers C3 and C2 (when I1 is in place) are usually sufficiently large to prevent this interference. If necessary, another isolator can be inserted in the path with R2.

The similarity of Fig. 1 to a phase lock synchronizer circuit continues into the IF components. The auxiliary IF signal from M1 is amplified in A1 and mixed with the output of a crystal controlled reference oscillator f_{B1} in mixer M3. In the phase lock case, however, A1 would be narrowband and would include amplitude limiting. M3 would be a phase detector and the output of a DC coupled low frequency amplifier A3 would be fed through a phase compensation network to the LO power supply to complete the feedback loop. The signal output would be taken directly from A2.

In the present feed-forward scheme, the bandwidths

of amplifiers A1, A2, and A3 and bandpass filter BPF1 must be large enough to handle all the frequency variation $\delta - \epsilon$. If A1 does not have limiting, then attenuator R4 is adjusted to provide the proper injection level for M3. A frequency counter coupled to the output of A1 is useful for setting and monitoring the offset $f_i + \delta - \epsilon$ to be sure that $f_i + f_{B1} + \delta - \epsilon$ remains well within the passband of BPF1.

On the other hand, the bandwidth of BPF1 must be small enough to prevent the leakage of f_{B1} through M3, BPF1, and M4 from interfering with the signal leaving M4. Use of double balanced mixers for M3 and M4 helps to minimize this leakage, and the balance of M3 also reduced the leakage of $f_i + \delta - \epsilon$ into M4. This latter frequency when mixed with $f_i + f_{B1} + \delta - \epsilon$ would also interfere with the signal leaving M4. If BPF1 attenuates these undesired leakages sufficiently, it ordinarily will also sufficiently attenuate the undesired sideband at $|f_i + \delta - \epsilon - f_{B1}|$ that is generated in M3.

Virtually all frequency jitter and drift in the signal is canceled out in mixer M4. The unshifted ($f_p = 0$) signal appears at exactly f_{B1} . Hence the passband of filter BPF2 can be made very small in interferometer and transmission measurements to reduce noise. RF image rejection in doppler shift measurements can also be accomplished conveniently in BPF2. If lower sideband ($f_p < 0$) signals and noise are to be rejected, BPF2 can be made with a sharp low frequency cutoff near f_{B1} . The image rejection achieved in this manner is much sharper than could be achieved with an RF filter in front of M2.

The basic circuit concept shown in Fig. 1 has been mentioned previously, but was regarded as impractical since only narrowband amplifiers (A1 and A2) were envisioned.⁸ Now, however, very low noise wideband transistor amplifiers are commercially available.⁹ Another circuit similar to the IF portion of Fig. 1 has also been used for measuring oscillator AM noise in the presence of large FM noise.¹⁰

Since both f_{B1} and the bandwidth of BPF1 must be large to accommodate large frequency drift, a third mixing stage may be needed to translate the signal down to a low frequency for signal processing. Such a stage is shown by the "optional" box in Fig. 1. Since there is negligible drift in f_{B1} , both f_{B2} and the bandwidth of BPF3 can be small.

In doppler-shift measurements, the signal processing may sometimes be simpler if the unshifted ($f_p = 0$) signal appears at DC. This effect can be produced by mixing the signal directly with the reference (either f_{B1} or f_{B2}) prior to signal processing. The result is similar to what would happen with simple homodyne RF mixing. The dynamic range should be considerably greater, however, because the RF oscillator AM noise and the $1/f$ noise in the RF mixer M2 should be much less at f_i than near DC. Furthermore, BPF2 or BPF4 can be used to provide rejection of $f_p < 0$ or $f_p > 0$ signals ("image rejection"). Since f_{B2} can be a much lower frequency than f_{B1} , image rejection of signals with low $|f_p|$ is easier in BPF4 than in BPF2.

APPLICATIONS

Table 1 summarizes some present and possibly future applications of the scheme of Fig. 1. Interferometer A has been installed on the PDX tokamak plasma machine at the Princeton Plasma Physics Laboratory. After an initial warmup of about one hour, the frequency difference of two water cooled klystron oscillators remains within the 20 MHz passband of BPF1 for several hours. The dynamic range (including waveguide and free space propagation loss through PDX) is at least 80 dB. The 30 MHz signal leaving BPF2 is sent to a phase lock receiver with a 100 kHz bandwidth. Both the reference and signal are then translated to 1 MHz and applied to a digital multifringe phase detector.¹¹

Without increasing the total number of frequency translation stages, the passband of BPF1 can be increased more than tenfold using the scheme denoted Interferometer B. Here the frequency is translated directly from 500 MHz to 1 MHz without the 30 MHz intermediate stage. Upgrade of the interferometer from the A to the B version would allow use of the interferometer without warmup or retuning.

An eight channel doppler-shift scattering system is also being built for PDX with the parameters listed in Table 1. By setting f_{B2} at 50 MHz, scattered frequency shifts out to both +60 MHz and -50 MHz can be simultaneously on a 110 MHz spectrum analyzer (see Fig. 2). Frequency shifts of this order may be encountered by scattering from waves associated with the ion cyclotron resonance. Drift waves, on the other

hand, generally produce shifts less than 1 MHz in magnitude.¹² Setting f_{B2} as low as 9 MHz and making a low frequency cutoff of 499 MHz in BPF2 still allows positive and negative doppler shifts f_p to be seen simultaneously and unambiguously out to $|f_p| = 1$ MHz. At the same time, positive f_p out to 101 MHz can still be detected on the 110 MHz spectrum analyzer. If only the drift waves are of interest, 9 MHz is superior to 50 MHz for f_{B2} since the signal processing may be simpler at lower frequencies.

Even simpler signal processing is possible at the expense of losing information as to the sign of f_p for $|f_p| < 1$ MHz, if the signal from BPF2 is mixed directly with the 500 MHz reference f_{B1} . There is still no ambiguity in sign for $|f_p| > 1$ MHz, however, since frequencies corresponding to $f_p < -1$ MHz are rejected by BPF2 when that filter has a 499 MHz low frequency cutoff.

Finally, Table 1 shows possible parameters for a swept or fixed frequency transmission measurement. In a swept measurement, both backward wave oscillators (BWOs) would probably require a leveling loop to ensure reasonable measurement accuracy. A 2-4 GHz bandwidth for A1 and A2 allows the scheme to work as long as the frequencies of the two BWOs track within 2 GHz over the entire sweeping range. Examination of typical BWO frequency versus tuning voltage curves shows that such tracking can indeed be accomplished over greater than 30 GHz ranges near 100 GHz.

Since microwave filters with percentage bandwidths as

small as about 0.1% are readily obtainable, the system noise bandwidth can be as small as about 1 MHz when a 1200 MHz center frequency is used for BPF2. By introducing another frequency translation stage such as shown in the "optional" box in Fig. 1, the final noise bandwidth might be reduced to a few kilohertz even with megahertz of BWO FM noise and drift.

RESIDUAL NOISE

In practice, the minimum noise bandwidth of the scheme in Fig. 1 is probably limited by the propagation delay time through the measurement path. If this delay is more than several nanoseconds, FM noise modulation frequencies in the tens of MHz, such as are common on BWOs, will cause $\delta\phi$ coming from mixer M1 to differ somewhat from that coming from mixer M2. The signal leaving M4 will then have some residual frequency jitter. (In an interferometer, the closely related residual phase jitter is more significant.)

More quantitatively, let the FM noise spectrum of the transmitter be described in terms of an r.m.s. deviation Δf as a function of f_m , the modulation frequency. The square of the function $\Delta f(f_m)$ has units of $(\text{Hz})^2$ per Hz. The total r.m.s. linewidth $(\delta f)_1$ of the oscillator can be represented as

$$(\delta f)_1^2 = \int_{f_c}^{\infty} [\Delta f(f_m)]^2 df_m, \quad (1)$$

where f_c is some arbitrary low frequency cutoff separating "long term drift" from "FM noise."

In the present cracking circuit, the frequency jitter is suppressed at all modulation frequencies f_m much smaller than the reciprocal of the differential delay τ for the reference and measurement paths to mixer M4. Specifically, the r.m.s. frequency jitter $(\delta f)_2$ remaining in the signal leaving M4 will be

$$(\delta f)_2^2 = \int_{f_c}^{\infty} \left[\Delta f(f_m) \right]^2 (\sin 2\pi f_m \tau)^2 df_m. \quad (2)$$

Notice that the FM noise spectrum of the local oscillator is unimportant for the scheme in Fig. 1, unless BPF1 has such a narrow bandwidth that it introduces significant delay in the path from M1 to M4 relative to the path from M2 to M4.

An equation such as (2) also applies in cases where the two high frequency oscillators are synchronized by a feedback loop (AFC or phase lock). The residual jitter $(\delta f)_2^2$ then depends, however, on the local oscillator FM noise as well as the transmitter FM noise. Except for the low frequencies f_m within the loop bandwidth, which are unimportant in (2), the FM noise of the synchronized oscillator will be the same as if it were free running. If the transmitter and LO have similar noise spectra, then the total jitter $(\delta f)_2^2$ will be about double that for the circuit in Fig. 1.

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TABLE I APPLICATION TO MILLIMETER-WAVE MEASUREMENTS

MEASUREMENT	OSCILLATOR TYPES	f_o (GHz)	f_{i+1} (MHz)	f_{B1} (MHz)	BPFl (MHz)	f_p (MHz)	BPF2 (MHz)	f_{B2} (MHz)
INTERFEROMETER A	klystrons	140	150-170	30	180-200	---	29.8-30.2	---
INTERFEROMETER B	klystrons	140	150-450	500	650-950	---	490-510	1
SCATTERING	EIO+klystron	140	50-450	500	550-950	0.01-1 0.01-60	490-610 450-560	9 50
SWEPT TRANSMISSION	BWos	75-115	2000-4000	1200	3200-5200	---	1199-1201	---

EIO = EXTENDED INTERACTION OSCILLATOR

BWO = BACKWARD WAVE OSCILLATOR

Fig. 1. Superheterodyne Tracking Circuit.

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 Fig. 2. An example of frequencies present in the scattering system of Table I. ($\epsilon = 400$ MHz, $\delta = 500$ MHz, $f_i = 250$ MHz, $f_p = 30$ MHz)