Quadrature sampling phase detection

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A fast sampling demodulation method for phase modulated carriers is described. A reference wave is used to generate clock pulses for a digitizing instrument that measures directly the in-phase and quadrature components of the modulation. Application to demodulation of signals from a multi-channel phase-sensitive polarimeter is discussed.

I. INTRODUCTION

The demodulation of phase encoded high frequency carrier signals is a problem encountered in many applications and areas of research. Numerous electronic methods for demodulation of the carrier, including digital counting techniques and quadrature mixing demodulation have become standard.[1] Alternatively, the waveforms can be digitized and the phase extracted computationally.[2,3] As the carrier frequency increases however, this approach can place severe demands on digitizer speed and memory. This is especially true when more than a single carrier is to be recorded. We here report a simple sampling quadrature method for phase demodulation that is optimum in the sense that the digitizer needs only sample at twice the information bandwidth rather than at twice the carrier frequency.

The technique uses interleaved in-phase and quadrature clock pulses at twice the carrier frequency $f=\omega/2\pi$ (or a subharmonic of f) as an external clock for digitization of the phase modulated carrier. This clock is derived from an unmodulated reference frequency signal using a tracking phase-locked loop (PLL) and binary counter. As shown in Sec. II, the resultant digitized signals are proportional to the cosine and sine of the phase modulation $\varphi(t)$ which can then be extracted by the computer. The phase noise is limited by the digitizer discretization error and noise on the carrier. It is insensitive to changes in carrier frequency (determined by the tracking range of the PLL) and can be compensated for clock jitter. As well, the carrier amplitude, weakly time-varying dc offsets, and even harmonic contamination can be recovered and/or removed. Many channels can be recorded using the same external clock, obviating the need for electronic phase comparators. This is a great advantage in multi-channel systems where many phase detectors must be fabricated, calibrated, and maintained through the life of the experiment.[4,5]

In high frequency appplications, it may not be possible or desirable to use a reference clock at $4f$. Indeed, the digitization rate needs to be sufficient only to capture the temporal variation in the phase modulation $\varphi\ (t)$. As shown in Sec. III, it is straightforward, using binary counters, to produce a clock composed of a subharmonic of f and its quadrature partner from the PLL output.

A possible drawback of this sampling method is that it may be necessary to digitize a signal of known fixed frequency in order to calibrate the timebase. This is because the clock rate depends on the instantaneous carrier frequency which can be time dependent. This is the case for the scanning far-infrared laser interferometer/polarimeter application discussed in Sec. IV. This instrument is being developed for measurement of plasma density in the H1 heliac (helical axis stellarator) at the Australian National University.[6]

Though the sampling rate need be sufficient only to accommodate the information bandwidth the digitizer must be able to track and hold accurately signals at the carrier frequency rate. The feasibility of the method is thus limited by present digitizer technology to the range $f \lesssim 10$ MHz for 12 bit dynamic range.

II. METHOD

We consider a phase modulated signal $S=A(t)\cos[2\pi f(t)t+\varphi(t)]$ and its unmodulated reference frequency counterpart. For simplicity, we ignore the amplitude variations but allow the instantaneous frequency f to be time dependent. Using a frequency domain approach, we later indicate how the analysis can allow for time-dependent dc and higher harmonic components.

We represent the clock signal derived from the reference wave by the idealized sampling comb

$$R(t)=\sum_{n=0}^{\infty} \delta\left[t-\left(\frac{2M+1}{4}\right)n\tau(t)\right], \tag{2.1}$$

where $\tau=1/f$ is the wave period, M is a positive integer, and the sampling commences at $t=0$. (The effects of time offsets and sampling jitter are discussed in Sec. III.) The case $M=0$ corresponds to sampling at the maximum rate $4f$. For M even, the digitized signals are given respectively by the repeated time sequence

$$\begin{aligned} S_n &= \{\cos[\varphi(nT)],\ -\sin[\varphi((n+1)T)],\\ &\quad -\cos\varphi[(n+2)T],\ \sin\varphi[(n+3)T]\}\\ &\equiv\{s_{0n}, s_{1n}, s_{2n}, s_{3n}\}, \quad n=0,4,8\ldots \end{aligned}$$

where $T=(2M+1)\tau/4$ is the sampling interval. Though the samples are recorded at different times, it is possible to estimate simply the quadrature component of, say, the sample s_{1n} at time $(n+1)T$, by linearly interpolating between the components s_{0n} and $-s_{2n}$. The arctangent of the in-phase and quadrature samples then yields the instantaneous value of the phase modulation.

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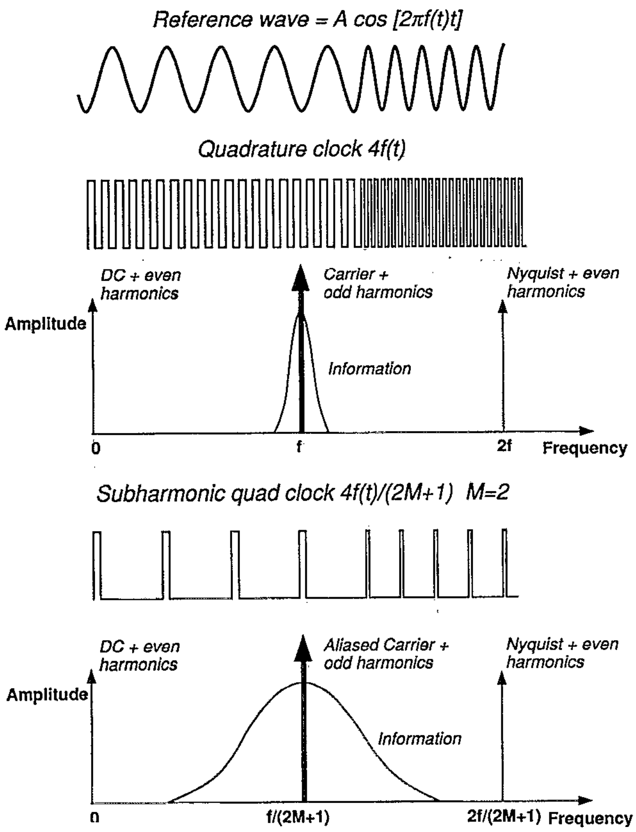

FIG. 1. Pictorial representation illustrating the mapping of time domain information into the frequency domain. When a subharmonic clock is used to digitize the signal carrier, the information bandwidth is increased by a factor $(4M+1)$.

It is instructive to examine the frequency domain representation of the $4f$ sampling scheme. Because the sampling rate varies with the instantaneous carrier frequency, the carrier phase is sampled at a constant rate. The discrete Fourier transform of the data samples thus covers a fixed interval in the domain of the phase transform. The carrier and odd harmonics are mapped to the center of this passband and dc and even harmonics are aliased to zero and half the sampling frequency. This allows for filtering and elimination of unwanted even harmonic noise. Should odd harmonic contamination be troublesome, an $8f$ scheme would facilitate its removal. The information is assumed to occupy some region within $\pm\Delta f/2$ of the carrier as illustrated in Fig. 1. For subharmonic sampling ($M>0$), the fraction of the passband occupied by the phase information will be greater by a factor $2M+1$, it being ultimately aliased back into the passband for large M. We note that frequency domain post-processing techniques allow the trigonometric quantities to be filtered, interpolated, and compared on a common time grid.

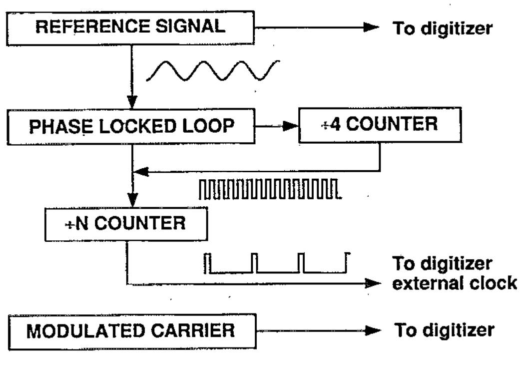

FIG. 2. Block diagram showing the implementation of the sampling phase detector.

III. IMPLEMENTATION

The block diagram for the apparatus is shown in Fig. 2. The required $4f$ clock is obtained using a PLL (NE 465) and divide-by-four counter. An additional counter selects the desired subset of quadrature pulses for use as the external clock to the digitizer. In effect, the circuit is a standard frequency synthesizer.[1]

The time constant for the PLL internal phase discriminator is a compromise between minimizing jitter in the zero-crossing times for the output signal at $4f$ and minimizing the acquisition time for sudden changes in carrier frequency. The PLL also introduces a frequency independent time delay τ_0. These sources of phase error are easily compensated by recording the reference wave (from which the clock was generated) as well as the signal wave. Provided the digitizer records the signals simultaneously, the phase jitter and offsets are common and can be subtracted.

An additional identical PLL stage can be installed for the phase modulated signal. The reference and signal upshifted outputs are then directed to a flip-flop and low pass filter to remove the high-frequency component. The result is an analog measure of the phase difference that can be compared with that obtained using the digitizing method. This is a standard phase detection method except that the phase reset range (normally π radians) is, by virtue of the frequency upshift, reduced to $\pi/4$ radians. However, this allows some practical increase in recoverable information bandwidth by allowing an increase in the final low pass filter bandwidth. Clearly, a frequency downshift will allow the reset range to be extended to an integral multiple of π radians. This can be obtained without loss of information bandwidth provided the carrier frequency is sufficiently great. This flexibility is an advantage when attempting to either match a small phase shift into a given voltage range or when trying to avoid reset ambiguities when the phase variation is large. As with other such electronic demodulation methods, however, the circuit performance can be compromised by mismatched PLL time delays, relative time jitter of the PLL output square waves and leakage of the final low pass filter. We compare the electronic phase comparator signals with the phase extracted using various sampling methods in the next section.

IV. EXPERIMENT

The sampling clock was tested using signals derived from a multi-channel far infrared phase-sensitive polarimeter.[3] Because of the Faraday effect, a linearly polarized probing wave traversing a magnetized plasma will suffer some polarization rotation θ. Using suitable post-processing

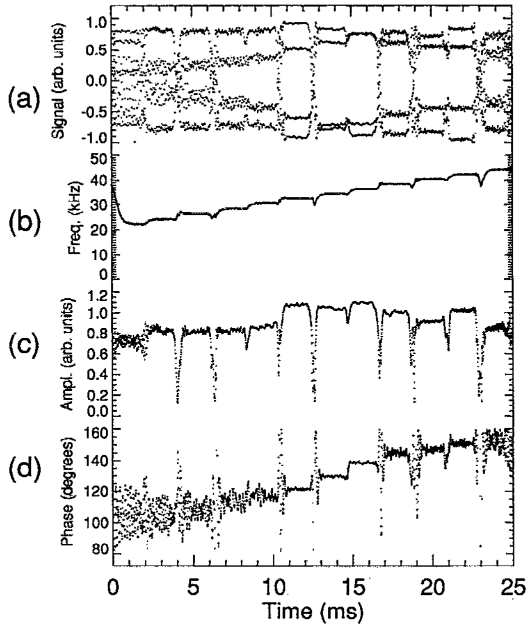

FIG. 3. (a) The quadrature sampled reference wave. There are four samples per wave cycle in each of the 12 distinct fringe bursts. (b) The carrier frequency as monitored by the calibrated VCO control signal. (c) The inferred reference signal amplitude. (d) The computed reference wave phase shift. See text for discussion.

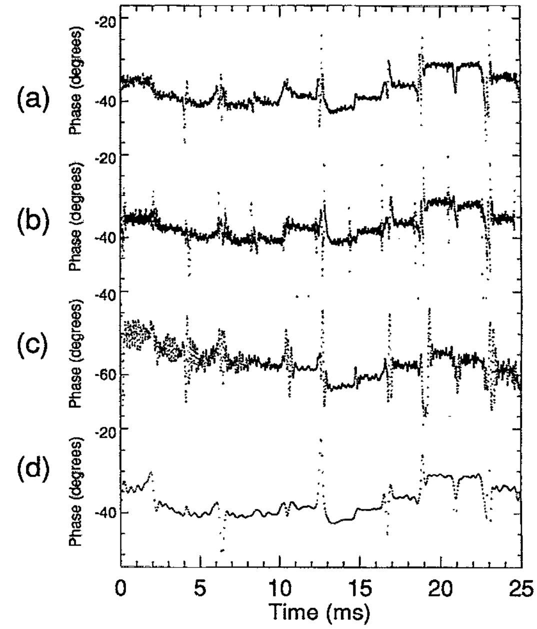

FIG. 4. (a) The phase difference between the polarimeter signals obtained using the quadrature sampling method. (b) The phase difference between the computed analytic signals. (c) The phase profile as measured electronically. (d) The phase difference for the $M=2$ quadrature clock. See text for discussion.

optics, this rotation can be registered as a relative phase shift 2θ between signals from detectors sensing orthogonal polarization components.

It is possible to scan the probing laser beam across the plasma region by reflecting it from the edge of a rapidly rotating disk grating whose grating constant varies discretely (or continuously) with disk rotation angle.[3,7] Because of the rotation, the beam is also Doppler shifted by an amount depending on the reflection angle from the grating. The intermediate-frequency (IF) phase-modulated carrier signal is generated by mixing local oscillator and Doppler shifted probing beams in a Schottky diode detector. The detector signal is thus composed of adjacent fringe bursts having discrete carrier frequencies f_n changing in step fashion over the duration of the grating revolution. For a continuously changing grating pattern, the carrier signal frequency $f(t)$ varies in a continuous fashion with time. The IF frequencies range over an octave in frequency and typically occupy the range 10–200 kHz. Accurate demodulation of such "broadband" signals can pose special problems for a standard electronic phase detector but are ideally suited to the new method.

The optical system used for the experiments reported below is described elsewhere.[3] To test the sampling method, one of the polarimeter detector outputs is taken as the phase reference and used to produce the quadrature clock for digitization of the second polarimeter signal. Also digitized were the reference wave itself, the VCO control voltage and a fixed frequency 5 kHz oscillation for timebase calibration. The quadrature clock is supplied to a Transiac 4012 CAMAC crate controller which hosts a set of Transiac 2825 4 channel 250 kHz digitizers. These units record simultaneously on all four channels with 12 bit resolution, an acquisition time of 1 μs and internal clock jitter of 100 ps.

A typical set of fringe bursts is shown in Fig. 3(a). The timebase has been corrected for the variable sampling rate. The calibrated VCO control voltage follows the changing carrier frequency which ranges between 20 and 50 kHz as seen in Fig. 3(b). The reference phase and amplitude determined from the set of quadrature samples are shown in Fig. 3(c) and 3(d) respectively. The signals at the transitions between channels can be noisy due to the decrease in signal amplitude. The remaining high level of noise present on some of the channels is due to half-harmonic and even harmonic contamination caused respectively by parasitic third order retro-reflected beams and double-pass second order beams in the return path from the grating. As well, the phase profile across the channels shows a linear variation with frequency equivalent to a fixed PLL time delay of $\sim 9 \mu$s.

The phase difference between the polarimeter signals digitized using the full bandwidth quadrature clock ($M=0$) is shown in Fig. 4(a). Note that clock jitter and time delays have been almost completely compensated by subtracting the phase profiles obtained separately for the reference and signal channels. The absolute value and channel to channel variation of the phase shift is of no significance here (see Ref. 3). There is some systematic variation in the phase profile within a given fringe burst. This is is evident upon comparison with the phase profile obtained by digitizing the two polarimetric signals at a fixed 200 kHz rate and computing the corresponding analytic signals, from which can be extracted the relative phase as shown in Fig. 4(b). This variation is reproducible from scan to scan and can be accurately compensated as described in (Ref. 3). Most of the remaining phase noise arises from amplitude noise on the carrier, the signal to noise ratio in these experiments being about 100. The effects of discretization error are negligible. Of course the noise can be reduced by decreasing the bandwidth as described below.

The phase shift recorded by the electronic phase detector described in Sec. III is shown in Fig. 4(c). A linear phase ramp corresponding to a mismatch of the PLL delays of 0.17 μs is apparent on the phase profiles. The high level of noise can be attributed to leakage of the low-pass filter to the half-harmonic contamination mentioned above. The delay mismatch is too small to account for such phase errors at these frequencies. These problems are easily removed when the waveforms are directly sampled.

The effect of increasing the sampling period ($M=2$) is shown in Fig. 4(d). Prior to phase extraction, these more sparsely sampled signals (reference and phase) are filtered using a fixed cosine window centered on the carrier and with full width equal to half the passband. Since the sampling rate is reduced fivefold, the information bandwidth is effectively increased by the same factor (see Fig. 2) with the result that the recovered phase bandwidth is reduced by an order of magnitude compared with Fig. 4(a) where no filter was applied. Note, however, that the systematic variations persist and that the absolute phase is faithfully recovered.

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