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GROUP-DELAY ERRORS DUE TO COHERENT INTERFERENCE

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ABSTRACT

Time and frequency transfer techniques that rely on spread spectrum group-delay measurements are prone to potentially large timing errors caused by coherent interference that is time delayed. Reducing the level of the coherent interference, increasing the pseudo-random code rate, or using a fractional correlator can minimize these timing errors.

1. INTRODUCTION

The time-and-frequency community uses spread spectrum communication links for the transfer of time and frequency information from laboratory to laboratory or more commonly, the dissemination of time and frequency to users across the world. The two systems that are widely used for the transfer of time-and-frequency information are the global positioning system (GPS) and two-way satellite time-and-frequency transfer (TWSTFT). In both of these systems a timing signal is encoded on a carrier signal using bi-phase spread spectrum modulation. Data taken over the past ten years indicate that although TWSTFT is a two-way system where there should be common mode rejection of path delays, the delay stability of this method of transfer has generally not been much better than common view GPS. However, there is a recent TWSTFT link that shows a factor of 5 improvement in stability over other TWSTFT links [1]. In this paper we discuss limitations on time and frequency transfer due to fundamental limitations in group-delay measurements. Both GPS and TWSTFT rely on a group-delay measurement to transfer time and frequency information. Understanding group-delay measurements and their limitations are essential to establishing a stable method of time and frequency transfer. The simple group-delay theory in the case of a propagating sine wave is presented first and is followed by the group-delay theory of a spread spectrum signal with correlators as the detecting elements.

2. SIMPLE GROUP-DELAY THEORY

Let us assume a group-delay measurement in which we are trying to measure the propagation time of a signal from point A to point B through some non-

dispersive propagation medium. Consider using a sinusoidal signal described by

$$\sqrt{2P} \sin(\omega t), \quad (1)$$

where the power level is P , the frequency is ω and t is time. The phase, ϕ of this signal after a propagation time of τ , is given as

$$\phi = \omega\tau \quad (2)$$

where τ is the phase delay. The group-delay, τ_g , is defined by

$$\tau_g = \frac{\partial\phi}{\partial\omega}. \quad (3)$$

Since the medium is assumed to be non-dispersive, the group-delay is equal to the phase delay when there is no interfering signal. Consider the case where a signal of power P_m , with the same frequency ω , delayed by time τ_d is summed with the original signal. The phase of the interfering signal with respect to the main signal is θ described by

$$\theta = \omega\tau_d, \quad (4)$$

where τ_d is the time delay between the main and interfering signal. Figure [1] shows that the combined

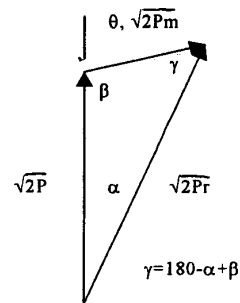


Figure 1. Vector diagram showing the relation between the main signal with power P , the delayed signal with power P_m , and the resultant vector sum with power P_r .

signal has an angular deviation α and a new level P_r . To keep the math simple we will make the assumption that the delayed signal is small compared to the main signal, so that we can use the small angle approximation

$$\sin(\alpha) \approx \alpha. \quad (5)$$

Starting with

$$\sin(\alpha) = \sqrt{\frac{P_m}{P}} \sin(\gamma), \quad (6)$$

we get, after using some trigonometric identities and the small-angle approximation, an expression for α described by

$$\alpha \approx -\sqrt{\frac{P_m}{P}} \sin(\omega\tau_d). \quad (7)$$

This expression for α is used to obtain the group-delay variation,

$$\Delta\tau_g \approx \frac{\partial\alpha}{\partial\omega} \approx -\sqrt{\frac{P_m}{P}} \tau_d \cos(\omega\tau_d). \quad (8)$$

$\Delta\tau_g$ is the difference between the phase delay and the group-delay when there is time delayed coherent interference. For example, if the group-delay were being measured to try to determine the electrical delay of a cable, $\Delta\tau_g$ would be the time error due to VSWR in the cable being measured. The group-delay variation has a peak value that varies linearly with τ_d , and goes through a full cycle when the phase between the main

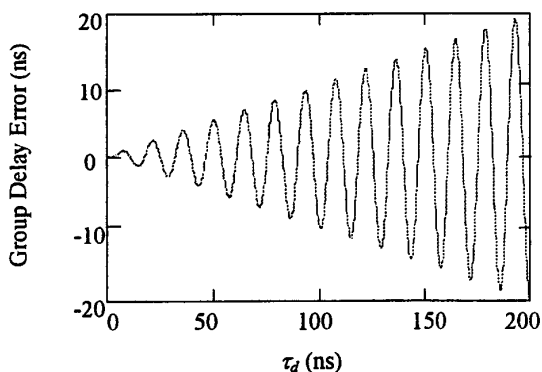


Figure 2. The group delay error $\Delta\tau_g$ as a function of the delay between the main and interfering signal τ_d for the an interfering signal level -20 dB from the main signal.

signal and the delayed signal goes through 2π . The

period of $\Delta\tau_g$ is equal to the period of the carrier frequency ω . The conclusion is that a group-delay measurement may not be a good estimate of the time delay in the presence of delayed coherent interference. For large τ_d the peak group-delay diverges from the phase delay without bound even for small values of P_m . Figure 2 is a plot of $\Delta\tau_g$ for increasing delay τ_d between the main and interfering signal.

3. EARLY/LATE CODE CORRELATOR

Using a spread spectrum modulated signal to measure the group-delay will limit the group-delay variations due to coherent interference. The group-delay deviation from the phase delay goes to zero as the time delay τ_d , between the main signal and delayed signal becomes larger and the delayed signal is no longer correlated with the main signal. An example of a system that uses a spread spectrum signal to measure time delays is GPS. For GPS, the delayed coherent interference could be due to multipath effects. Multipath signals can be large, in some cases the multipath signals may be only 10 dB below the direct signal. Multipath effects are well documented and the following derivations can be found in many publications [2,3]. The results can be generalized to cover delayed coherent interference from any source such as: multipath, reflections due to impedance mismatch in long antenna cables, multiple reflections through filters, near-field reflections in antennas, and reflections in waveguides. A simplified model of a noncoherent early/late (E/L) correlator such as is used in both GPS and TWSTFT systems is described below. The complete derivations of these equations are

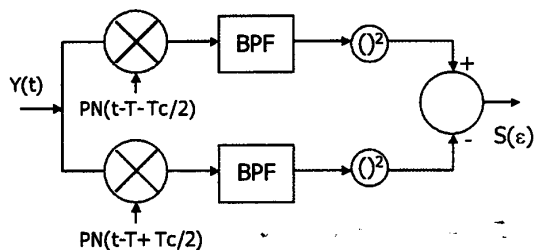


Figure 3. Code loop detector model.

presented in [2,3]. In Figure 3, the E/L code loop detector is displayed in its simplest form. The input signal $Y(t)$ is split and is the input to the early and the late correlator channels. The reference pseudo-random code sequence (PN) to the early and late correlators is spaced by a chip. The time variable is t , the reference sequence time estimate is T , and the chipping period is T_c . In both the early and late correlation channels, the output of the correlator is band-pass filtered and

squared. The code detector error signal $S(\epsilon)$ is the difference of the early and late correlation channel outputs. The error signal $S(\epsilon)$ is driven to 0 by the delay locked loop (DLL) in normal tracking operation. The tracking-loop error (t-T) is given by ϵ .

4. MEASURING GROUP-DELAY WITH A CORRELATOR

First we consider a group-delay measurement without an interfering signal. The spread-spectrum signal is propagating from A to B in a non-dispersive medium. The spread spectrum signal is

$$\sqrt{2P} \cdot PN(t) \cdot \sin(\omega \cdot t) \quad (9)$$

where P is the power level of the signal, $PN(t)$ is the pseudo-random code sequence and ω is the carrier frequency. After a propagation time τ , the signal

$$Y(t) = \sqrt{2P} \cdot PN(t - \tau) \cdot \sin(\omega \cdot t - \omega\tau), \quad (10)$$

is received and decorrelated by the E/L correlator. The error signal $S(\epsilon)$ is given by

$$S(\epsilon) = P \cdot R_{PN}^2(\epsilon - \frac{T_c}{2}) - P \cdot R_{PN}^2(\epsilon + \frac{T_c}{2}), \quad (11)$$

where $R_{PN}(t)$ is the auto-correlation function of $PN(t)$ and ϵ is defined as the code timing error. The code timing error is

$$\epsilon = \tau_g - \tau, \quad (12)$$

where τ_g is the measured group-delay and τ is the propagation delay as defined above. Figure 4 shows the

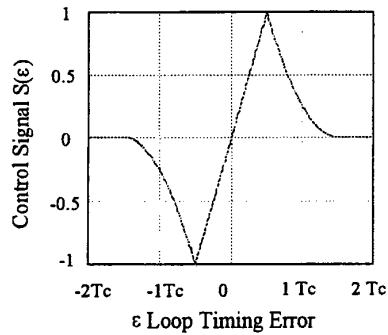


Figure 4. $S(\epsilon)$ without interference.

code detector error signal for the case of no interference. When the delay-locked loop is locked the function $S(\epsilon)$ is equal to 0

$$P \cdot R_{PN}^2(\tau_g - \tau - \frac{T_c}{2}) = P \cdot R_{PN}^2(\tau_g - \tau + \frac{T_c}{2}), \quad (13)$$

$$\tau_g = \tau, \quad (14)$$

and τ_g measures the desired quantity τ .

Now consider the case where there is a coherent interfering signal that is delayed by τ_d . The input signal $Y(t)$ is the sum of the direct signal and one delayed interfering signal:

$$Y(t) = \sqrt{2P} \cdot PN(t - \tau) \cdot \sin(\omega \cdot t + \omega\tau) + \sqrt{2P_m} \cdot PN(t - \tau - \tau_d) \cdot \sin(\omega \cdot t + \omega\tau + \theta), \quad (15)$$

where τ is the propagation time, ω is the direct signal carrier frequency and τ_d is the time delay of the reflected interfering signal with respect to the direct signal. θ is the phase of the interfering signal with respect to the main signal and is related to τ_d by equation (4). The error signal from the code loop detector $S(\epsilon)$ is

$$S(\epsilon) = P \cdot R_{PN}^2(\epsilon - \frac{T_c}{2}) + P_m \cdot R_{PN}^2(\epsilon + \tau_d - \frac{T_c}{2}) + 2P \cdot R_{PN}(\epsilon - \frac{T_c}{2}) \cdot P_m R_{PN}(\epsilon + \tau_d - \frac{T_c}{2}) \cdot \cos(\omega\tau_d) - P \cdot R_{PN}^2(\epsilon + \frac{T_c}{2}) - P_m \cdot R_{PN}^2(\epsilon + \tau_d + \frac{T_c}{2}) - 2P \cdot R_{PN}(\epsilon + \frac{T_c}{2}) \cdot P_m R_{PN}(\epsilon + \tau_d + \frac{T_c}{2}) \cdot \cos(\omega\tau_d). \quad (16)$$

The error signal is a function of τ_d , the time delay between the direct and reflected signal and P_m/P the ratio of reflected signal power to direct signal power. Figure 5 shows the code detector error signal for the

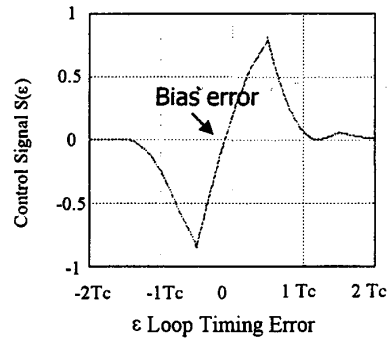


Figure 5. $S(\epsilon)$ with interference.

case where interference is present.

During tracking mode $S(\epsilon)$ is 0 and ϵ is not 0 as in the case where there is no interference. The group-delay error $\Delta\tau_g$ is obtained by solving for ϵ with the condition $S(\epsilon)=0$. $\Delta\tau_g$ is a complicated function of P , P_m , ω and τ_d . For $\tau_d < 0.5 T_c$, the expression for $\Delta\tau_g$ is manageable and is given by
For $P_m \ll P$, and when $\tau_d < 0.5 T_c$, Equation (17)

$$\Delta\tau_g = -\tau_d \left[\frac{P_m + \sqrt{PP_m} \cos(\omega\tau_d)}{P + P_m + 2\sqrt{PP_m} \cos(\omega\tau_d)} \right] \quad (17)$$

reduces to

$$\Delta\tau_g \approx -\tau_d \left[\sqrt{\frac{P_m}{P}} \cos(\omega\tau_d) \right], \quad (18)$$

which is identical to the expression for the sine wave signal (see Equation 9). The function for the group-delay error changes when τ_d becomes larger than T_c . With a spread spectrum signal the interfering signal begins to lose correlation with the main signal as τ_d becomes larger than the chip spacing of the correlator (in this example $T_c = 400$ ns). A plot of the group-delay error as a τ_d varies from zero to $1.5 T_c$, is shown in Figure 6 for the case of an interfering signal that is -20 dB lower in power than the direct signal. The carrier

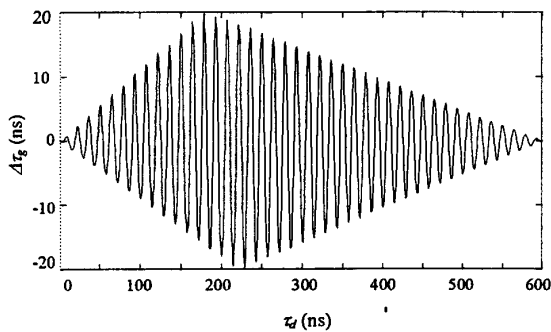


Figure 6. Group delay error as a function of τ_d . The chip period is 400ns and the detector is a 1/2chip E/L correlator.

frequency is 70 MHz and the chip rate is 2.5 MHz. The error becomes smaller for $\tau_d > 200$ ns. Understanding the group-delay error function can help us improve the spread spectrum time and frequency transfer systems. The following section discusses some possibilities.

5. MINIMIZING GROUP-DELAY ERRORS IN SPREAD SPECTRUM SYSTEMS

The most obvious method of reducing the group-delay error is by reducing the reflected signal power, P_m . Antennas can be positioned to reduce the effects of multipath. All impedances should be matched to minimize signal reflections in cables and system components such as filters and amplifiers. Phase stable cables should be used to minimize mapping small phase changes into large delay errors [4].

The other method is to exploit the correlation properties of the spread spectrum system. Recall that as τ_d becomes larger than half the chip spacing of the correlator the group-delay error begins to decrease. One method of decreasing the group-delay error is therefore to decrease the correlator spacing [3]. Decreasing the correlator spacing comes at an expense. The linear tracking region and the acquisition performance are degraded [3]. It may not be possible to operate under these conditions in cases where the signal-to-noise ratio is poor. Figure 7 shows the group-

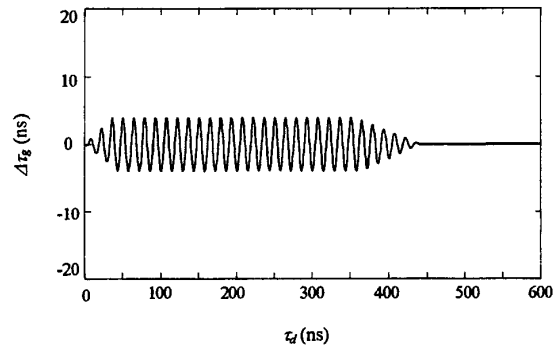


Figure 7. Group delay error as a function of τ_d . The chip period is 400 ns and the detector is a 1/10 chip E/L correlator.

delay error as a function of τ_d for the same signal conditions as in Figure 6 but with a one-tenth chip E/L correlator. In other words the reference code to the early and late correlators is spaced by one-tenth of a chip.

Another method of reducing the group-delay error is to use a higher chip rate. By increasing the chip rate you decrease the correlation time window during which the interfering signal affects the group-delay measurement. Figure 8 displays the group-delay error for the same signal levels as in Figure 6 but the spread spectrum signal is generated by a 20 MHz chip rate. The E/L correlator spacing is one-half of a chip.

In the TWSTFT system one of the sources for the group-delay error is the VSWR in the 30 m to 60 m

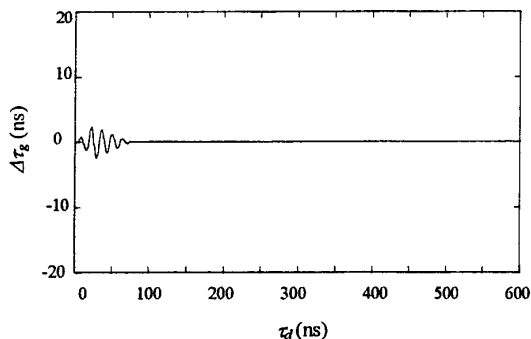


Figure 7. Group delay error as a function of τ_d . The chip period is 50 ns and the detector is a one-half chip E/L correlator.

cables that go from the laboratory to the earth station [4]. The cables lengths are such that reflected signals due to VSWR are delayed from the main signal by 300 to 400 ns. As has been illustrated in Figure 6, the group-delay error has the potential of being large. Making the reflected signal delay larger than 600 ns (greater than $1.5 T_c$) could make the group-delay error smaller without changing the chip rate of the modems. This could be accomplished by using an optic fiber RF link with a propagation delay greater than 600 ns. The delayed coherent interference caused by reflections at the ends of the fiber would be delayed by greater than 600 ns and would no longer be correlated with the direct signal. To implement this method of mitigating the group-delay error, optical fiber should be used and not a longer RF cable. Reflections in the RF cable can occur at random places along the cable due to dielectric imperfections, kinks, or temperature gradients. Reflections in optical fibers are, for the most part, constrained to occur only at the ends.

CONCLUSIONS

All group-delay measurements are sensitive to delayed coherent interference. Group-delay measurements using a swept sine wave have a group-delay error function that is proportional to the time delay of the interfering signal. The spread spectrum group-delay measurement systems have the advantage that the group-delay error function is 0 outside the correlator bandwidth. In addition, using the time correlation properties of spread spectrum systems can minimize the group-delay error function. Systems that suffer delay errors due to VSWR problems cannot be calibrated due to the sensitivity of the group-delay error

function on the phase of the carrier. The VSWR in ground station components is closely coupled to the environment and may be the major cause of instability in TWSTFT. New technology that is now available may be used to increase the stability of current methods of time and frequency transfer. Some of this technology includes optical fiber RF links, modems that are capable of chip rates in excess of 20 MHz, phase stable coaxial cables, and better antenna designs.

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