ENHANCEMENT OF LOW POWER AM SIGNAL IN AN ENVIRONMENT OF HIGH POWER AM INTERFERENCE[+]

تعزيز الاشارة المرغوبة الضعيفة ذات تضمين السعه في محيط يحوي على اشارة تداخل قوية مضمنة السعه ايضاً

Nabil. K. AL-Shamma [*]

Abstract :

A simple electronic signal to interference power ratio enhancement system incorporating phase – locked loop (PLL) is analyzed and constructed. The system is capable of producing accurate estimation of the interfering signal in an environment of high interference to the desired signal power ratio (ISR)i . The estimate is then used to cancel the interference with consequent enhancement of the desired weak signal. Theoretical analysis indicates that the output signal to interference power ratio is a function of the phase difference between the desired signal and the interference phase estimation. However it is found that at the worst case the power of the desired signal is comparable to the interference power enabling the desired signal to be heard.

المستخلص

تم تحليل وبناء منظومة الكترونية وذالك لتعزيز قدرة اشارة التضمين السعوي (AM) المرغوبة نسبة الى اشارة التداخل ذات القدرة العالية باستخدام دائرة قفل الطور (phase-locked loop)باستطاعة هذه المنظومة توليد تخمين دقيق لاشارة التداخل من ناحية الطور و التردد والسعة في محيط تكون فيه اشارة التداخل عالية القدرة مقارنة بالاشارة المرغوبة هذا التخمين سوف يتم طرحه من الاشارة المستلمة التي تحتوي على الاشارة المرغوبة و اشارة التداخل و بالنتيجه ستزداد قدرة الاشارة المرغوبه نسبة الى اشارة التداخل.

التحليل النظري بين ان نسبة اشارة التداخل المتبقية الى الاشارة المرغوبة تعتمد على فرق الطور بين الاشارة المرغوبة و اشارة التداخل . وبينت النتائج العملية عند استخدام المنظومة و الحصول على عملية القبض(lock) على اشارة التداخل تم (تعزيز قدرة الاشارة المرغوبة وتم سماع المحطة الضعيفة بصورة واصحة).

[+] Received on 1/8/2008 ,Accepted on 8/9/2009 .

[*] Asst . Lecturer , College of electronic and electrical technique

Introduction :

In the commercial AM spectrum, there are situation in which two stations coexist in the same AM channel, one station (the interference) is of high power, while the other (the desired) is of low power. In such a situation, the interference will be effective if it covers all the channel band of the desired signal. It is usual under such condition to estimate a signal corrupted by additive noise, by employing the adaptive interference canceling method, which requires for its operation a reference signal [1] [2] and [3].

A different approach is considered here. In essence it is a signal to interference power ratio enhancement technique that requires no such reference for its operation, since such signal is generated locally by the PLL. In the designed system. It is assumed that the interference is of higher power than the desired signal. This will facilitate the PLL operation to lock onto the interference while the desired signal is treated as noise. It is noteworthy to remember that the desired signal introduces some error in the estimated reference.

System analysis:

For two AM stations coexisting in the same RF channel, the received signal can be expressed as follows:

$$R(t) = A[1+\alpha\, i(t)] \cos(\omega_1 t+\Theta) + B[1+\beta m(t)] \cos(\omega_2 t+\phi) \ldots\ldots\ldots\ldots (1)$$

Where

A : a constant determining the power of the interference carrier ,

α : the modulation index , which is assumed to be less than unity ,

i(t) : the modulating interference normalized to have a maximum amplitude of unity ,

ω_1 : the carrier frequency of the interference ,

Θ: phase angle of the input interference signal ,

B : a constant determining the power of the desired carrier ,

β: the modulation index of the desired signal

m(t) : the desired modulating signal normalized to a maximum amplitude of unity ,

ω_2 : the carrier frequency of the desired signal , and in general $\omega_1 \neq \omega_2$, but they are in close proximity ,

ϕ :a random phase angle of the desired signal , uniformly distributed over the interval $(0 - 2\pi)$.

Now eqn . (1) can be written as follows :

$$R(t) = \sqrt{2S_i}\, \cos(\omega_1 t+\Theta) + \sqrt{2S_d}\, \cos(\omega_2 t+\phi) \ldots\ldots\ldots\ldots\ldots\ldots\ldots (2)$$

Where

$S_i = [A\,(1+\alpha\, i(t)]^2/2$ and $S_d = [B\,(1+\beta m(t)]^2/2$.

For the case of high power interference as compared with the desired signal , a narrow band circuit locking on the interfering frequency can produce an accurate estimate of the interfering signal .

The estimate can then be used to cancel the interference prior to demodulation . One such possible estimation technique is illustrated schematically in Fig . (1) . From this , it is seen that net result at the envelope detector input is expressed as follows :

$$A[1+\alpha\, i(t)] \cos(\omega_1 t+\Theta) + B[1+\beta m(t)] \cos(\omega_2 t+ \phi]$$

$$- A[1+\alpha\, i(t)]^{\wedge} \cos(\hat{\omega}_1 t+\hat{\Theta}) \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(3)$$

Where

$A[1+\alpha\, i(t)]^{\wedge}$, $\Theta^{\wedge}$ and $\hat{\omega}_1$ are the amplitude , phase and frequency estimate of the interference respectively , so that the effective interference residue , I_r , is given as :

$$I_r = A[1+\alpha\, i(t)] \cos(\omega_1 t+\Theta) - A[1+\alpha\, i(t)]^{\wedge} \cos(\omega^{\wedge}_1 t+\hat{\Theta})$$

$$= \sqrt{2S}_i \cos(\omega_1 t+\Theta) - \sqrt{2\hat{S}}_i \cos(\hat{\omega}_1 t+\hat{\Theta})\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(4)$$

By assuming a large enough power for the interference compared to that of the desired signal , the PLL will achieve lock and give perfect frequency estimation , i. e . $\omega_1 = \omega_1^{\wedge}$. With this assumption in mind and considering I_r as a sum of two sinusoids of the same frequency , I_r may be expressed by using the trigidentity (Appendix A) [4]
As :

$$I_r^2 = 2S_i + 2\hat{S}_i - 2\sqrt{2S}_i \sqrt{2\hat{S}}_i \cos(\Theta - \hat{\Theta}) \cos(\omega_1 t + \tau)\ldots\ldots\ldots\ldots\ldots\ldots(5)$$

The effectiveness of rejection of the interference signal is measured by the amplitude of the interference residue I_r , given as

$$I_r^2 = 2S_i + 2S^{\wedge}_i - 2\sqrt{2S}_i \sqrt{2\hat{S}}_i \cos(\Theta - \hat{\Theta}) \ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(6)$$

The power of the residue , S_r , is given by $S_r = I_r^2/2$.
When the PLL is locked on the interfering signal , the normalized output of the voltage controlled oscillator (VC0) [see Fig . (1)] can be expressed as $\sin(\omega_1 t+\hat{\Theta})$ where $\hat{\Theta}$ is the VC0 phase . The VC0 output is phase shifted by $\pi/2$, scaled by 2 to obtain $2\cos(\omega_1 t+\hat{\Theta})$, and multiplied by the received signal . The output of the multiplier is given as :

$$H(t) = 2A[1+\alpha\, i(t)] \cos(\omega_1 t+\Theta) \cos(\omega_1 t+\hat{\Theta}) +$$
$$2B[1+\beta m(t)] \cos(\omega_2 t+ \phi] \cos(\omega_1 t+\hat{\Theta})\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(7)$$

The output of the low pass filter (Bw) which must be designed to pass the modulating signals is given by :

$$F(t) = A[1+\alpha\, i(t)] \cos(\Theta-\hat{\Theta}) + B[1+\beta m(t)] \cos((\omega_2 - \omega_1) t + \phi - \hat{\Theta}))$$

$$= A[[1+\alpha\, i(t)] \cos(\epsilon) + B[1+\beta m(t)] \cos(\Delta\omega t+\Psi)$$

$$= \sqrt{2S}_i \cos(\epsilon) + \sqrt{2S}_d \cos(\Delta\omega t+\Psi)\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots(8)$$

Where

$\epsilon = \Theta - \Theta^{\wedge}$, $\Psi = \phi - \Theta^{\wedge}$ and $\Delta\omega = \omega_2 - \omega_1$.

For small value of ϵ, the first term in eqn. (8) then represents a good estimate of the interfering signal whereas the second term represents the "noise" in the amplitude estimate[5] , therefore

$F(t) = \sqrt{2S}_I{}^{\wedge} = \sqrt{2S}_i \cos(\epsilon) + \sqrt{2S}_d \cos(\Delta\omega t + \Psi)$

From eqns . (6) and (8) , one obtains

$I_r^2 = 2S_i + [\sqrt{2S}_i \cos(\epsilon) + \sqrt{2S}_d \cos(\Delta\omega t + \Psi)]^2 - 2\sqrt{2S}_i [\sqrt{2S}_i \cos(\epsilon) + \sqrt{2S}_d \cos(\Delta\omega t + \Psi)]$

$\cos(\epsilon)$

$= 2S_i + 2S_i \cos^2(\epsilon) + 2\sqrt{2S}_i \sqrt{2S}_d \cos(\epsilon) \cos(\Delta\omega t + \Psi) + 2S_d \cos^2(\Delta\omega t + \Psi) - 4 S_i \cos^2(\epsilon)$

$- 2 \sqrt{2S}_i \sqrt{2S}_d \cos(\epsilon) \cos(\Delta\omega t + \Psi)$

$I_r^2 = 2S_i - 2S_i \cos^2(\epsilon) + 2S_d \cos^2(\Delta\omega t + \Psi)$……………..……………(9)

Hence , by using the trigidentity $\cos^2(\epsilon) = [1+\cos(2\epsilon)]/2$, I_r^2 may be expressed as

$I_r^2 = S_i - S_i \cos(2\epsilon) + 2S_d \cos^2(\Delta\omega t + \Psi)$……………………………(10)

On the expansion of the cosine term as a power series , i . e .

$\cos(2\epsilon) = 1 - (2\epsilon)^2 / 2! + (2\epsilon)^4 / 4!$ …………..

and assuming $2\epsilon \ll 1$, as will be shown later , the cosine term can thus be expressed as : $\cos(2\epsilon) \approx 1 - 2\epsilon^2$ and eqn . (10) reduces to

$I_r^2 = 2S_i \epsilon^2 + 2S_d \cos^2(\Delta\omega t + \Psi)$……………..……………………(11)

Thus , the residue of the interference power at the input of the detector is expressed as :

$S = I_r^2/2 = S_i \epsilon^2 + S_d \cos^2(\Delta\omega t + \Psi)$…………………………………(12)

Reference [6] has previously shown that for a PLL with input composed of $\sqrt{2E} \cos(\omega t + \Theta)$ plus white noise of double-sided spectral density $\eta/2$ and loop bandwidth B_L , the phase error variance is given by :

$\sigma^2 = \eta B_L / E^2$

In the present case of interest , the PLL input is composed of $2\sqrt{2S}_I \cos(\omega_1 t + \Theta)$ plus the desired signal power S_d and bandwidth W which to the narrow band PLL appears like white noise of spectral density $S_d / 2W$. thus , the error variance may be expressed as :

$\sigma^2 = 2S_d B_L / 2WS_i = B_L S_d / WS_i$

Then, one may write [4]

$\epsilon^2 = \sigma^2 = B_L S_d / (W S_i)$………………………………………………(13)

It is noteworthy that the condition assumed above , i.e. high input interference to signal power ratio [S_i / S_d] and B_L / W being of the order of 1/40 * ,well justifies the for mentioned assumption where $2\epsilon << 1$.

Now by combining eqns .(12) and (13) , S_r may be expressed as :

$S_r = S_i (B_L \; S_d) / (W \; S_i) + S_d \cos^2(\Delta \omega t + \Psi)$

$\quad = S_d \; [B_L \; / W + \cos^2(\Delta \omega t + \Psi)]$

Therefore,

$S_r / S_d = [B_L + W \cos^2(\Delta \omega t + \Psi)] / W$ ………………………………(14)

It is of interest to note that under the assumed high $(ISR)_i$ condition the power residue in the interfering signal is independent of its received power ; for specified B_L and W , it depends only on the phase difference between the desired signal and the phase estimation of the interference . this phase is in turn a function of time .

On the other hand , for small input $(ISR)_i$ the PLL will not lock onto and track the interfering signal. Consequently the system will generate amplitude and phase estimates which are unrelated to the interference. Under this condition , the circuitry should be disabled .

Systemdesign:

The detailed block diagram of the system is shown in Fig . (1)

It consists basically of a second order PLL with low pass filter and correcting network . The free running frequency (fo) of the VCO is adjusted to a fixed value . The PLL parameters are given in table (1) shown below: [7][8] and [9].

Loop Gain rad . / sec .	Natural frequency rad . / sec .	Damping factor (ξ)	Loop bandwidth (B_L) Hz
50000	100	0.707	106

---

* In the present design B_L = 106 HZ and assuming W=4 KHZ .

B_L : Loop bandwidth of PLL.

The VCO signal and its $\pi/2$ phase shift are taken from an HP function generator.

The loop bandwidth need not be wide since the PLL is required to lock onto a constant frequency . This feature will improve the operation of the system as is clear from eqn .(14) .

The bandwidth of the low pass filter (Bw) was chosen as 10 KHz.It must be noted that it is not advisable to reduce this filter bandwidth because this may introduce some delay which in turn affect the subtraction process **.**

It is important to refer here that the multiply by 2 block which is the gain used by the theoretical analysis has to be adjusted manually in practice in order to obtain best estimate of the interference**.** This block is account for the overall gain of the system.

Results and conclusion:

The system designed and constructed in the course of this work has been tested in the laboratory. Two AM independent signals have been generated, one of high power which represents the interference, the other of low power which represents the desired signal. Fig.(2) shows the high power interference signal with and without the desired signal together with the estimated value of the interference amplitude at the output of the low-pass filter (the F-signal). Fig.(3) shows the H-signal which is the output of the multiplier with and without the desired signal. Here it is clear that the desired signal behave as a noise in the interference amplitude estimator.

The output of the multiplier is filtered by the low-pass filter Bw to obtain the F-signal which is as mentioned the amplitude estimate of the interference.

Also it is important to say here that it is difficult to measure practically the residue of the interference power at the input of the envelope detector for comparison with the theoretical value obtained from eqn.(14) because we cannot separate the residue of the interference from the desired signal since they occupy the same band. However , on applying the system and obtaining lock condition , a significant enhancement of the desired signal reception was obtained , and the station of interest was clearly heard .

The theoretically effective enhancement expected in eqn . (14) could not be achieved in practice due to implementation errors and "sloppiness" of the interfering signal . Furthermore, after enhancement of the desired signal , it did not exhibit any power fluctuation with respect to time as indicated in eqn . (14) . This may be attributed to difference in carrier that results in rapid periodic change of the desired signal amplitude with time which could not be recognized by the audiences.

Appendixa:

By assuming X and Y to be two sinusoidal signals having the same frequency , they may be then expressed as :

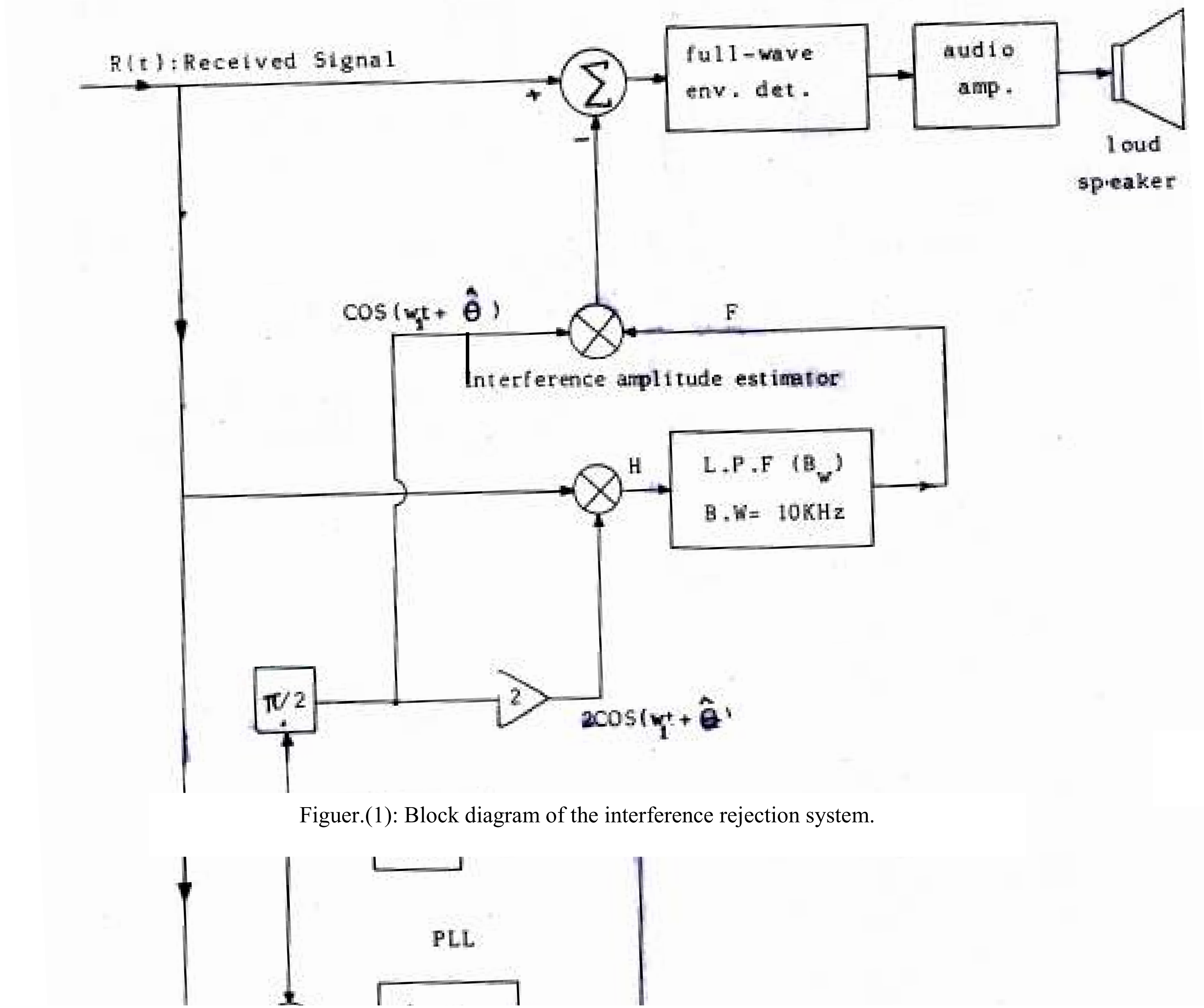
$X = A\cos(\omega t + a)$, $Y = B\cos(\omega t + b)$

Their sum Z can be expressed as :

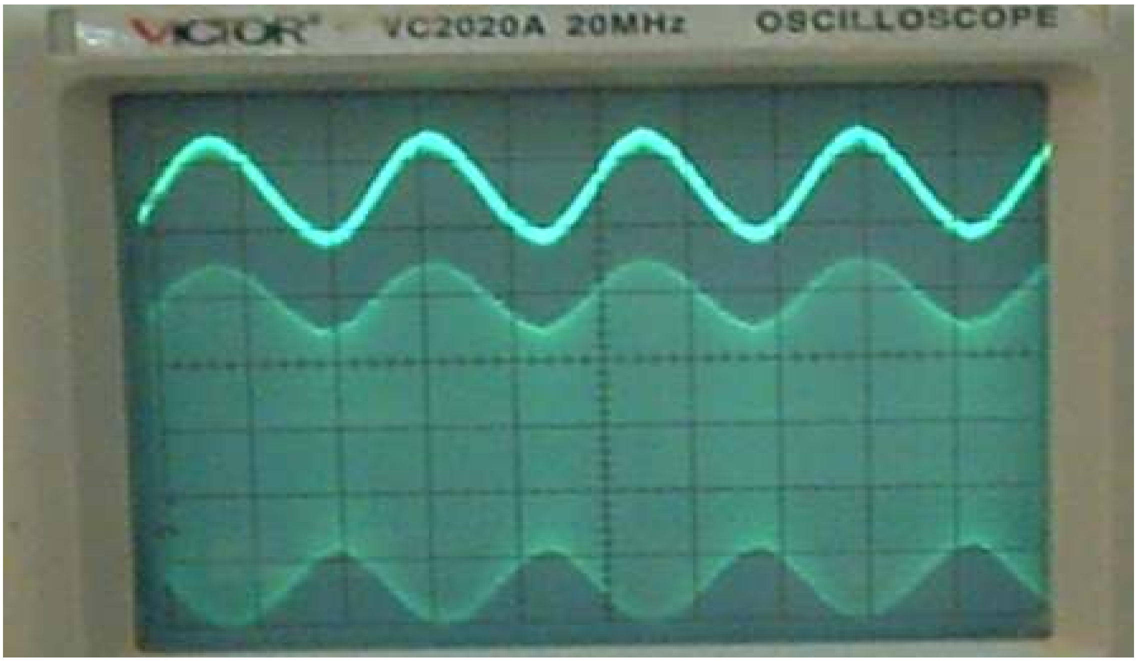
$Z = [A^2 + B^2 + 2AB\cos(a - b)]^{1/2}\cos(\omega t + \alpha)$

and

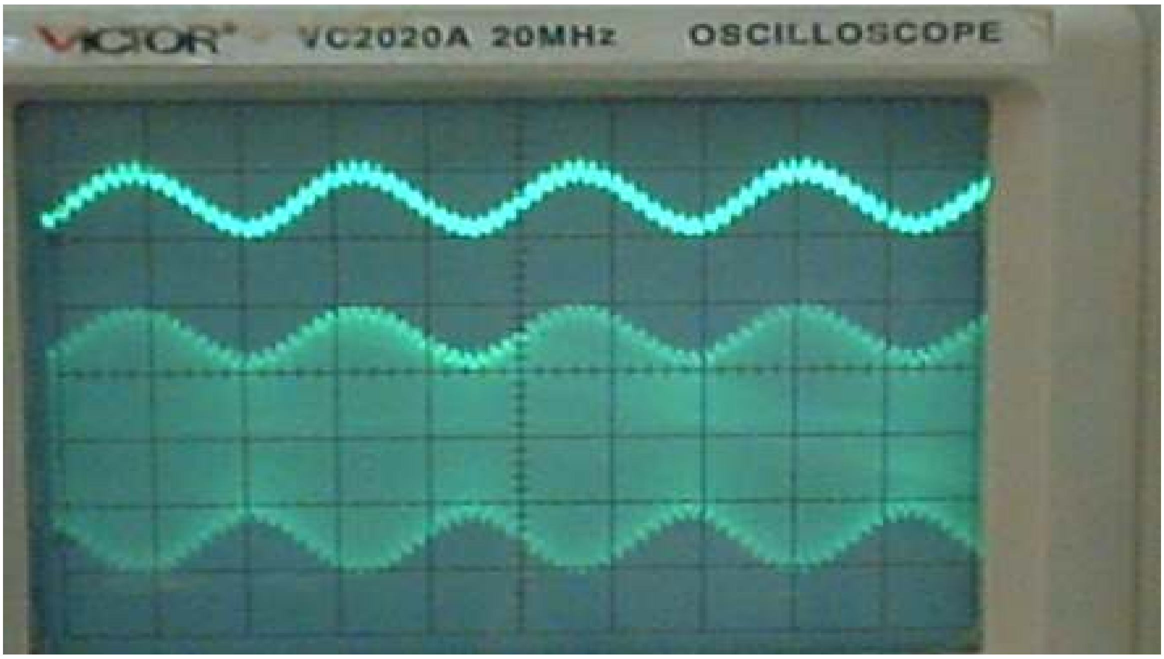
$\alpha = \tan^{-1}(A\sin(a) + B\sin(b)) / (A\cos(a) + B\cos(b))$



Figuer.(1): Block diagram of the interference rejection system.

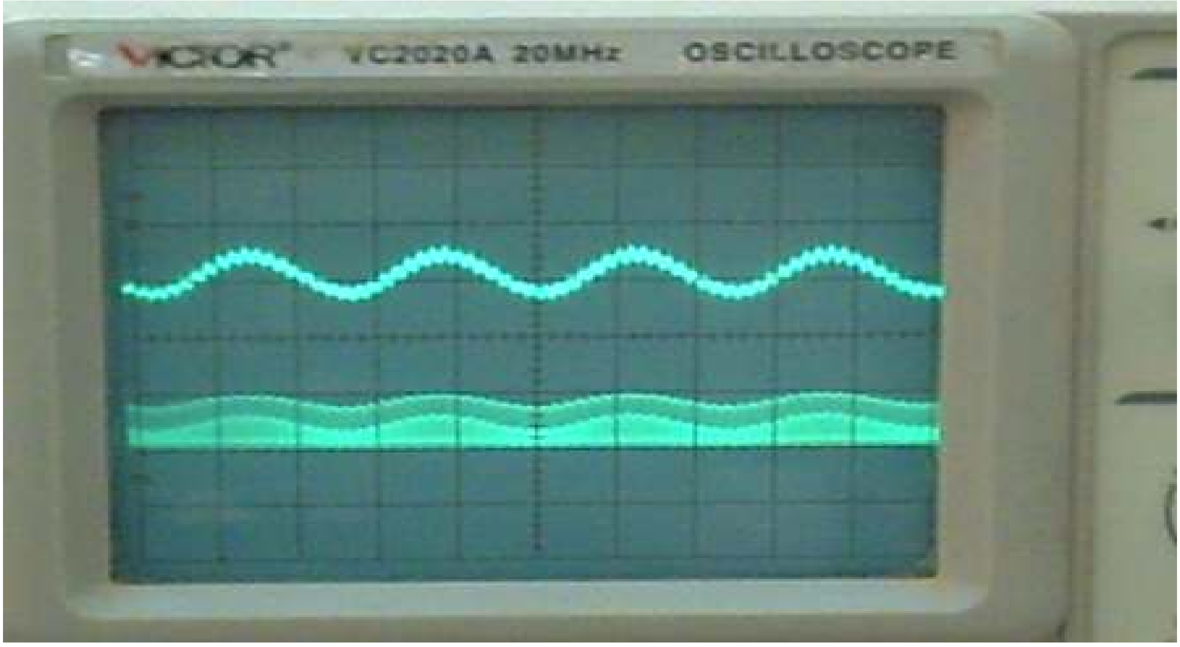


(a)

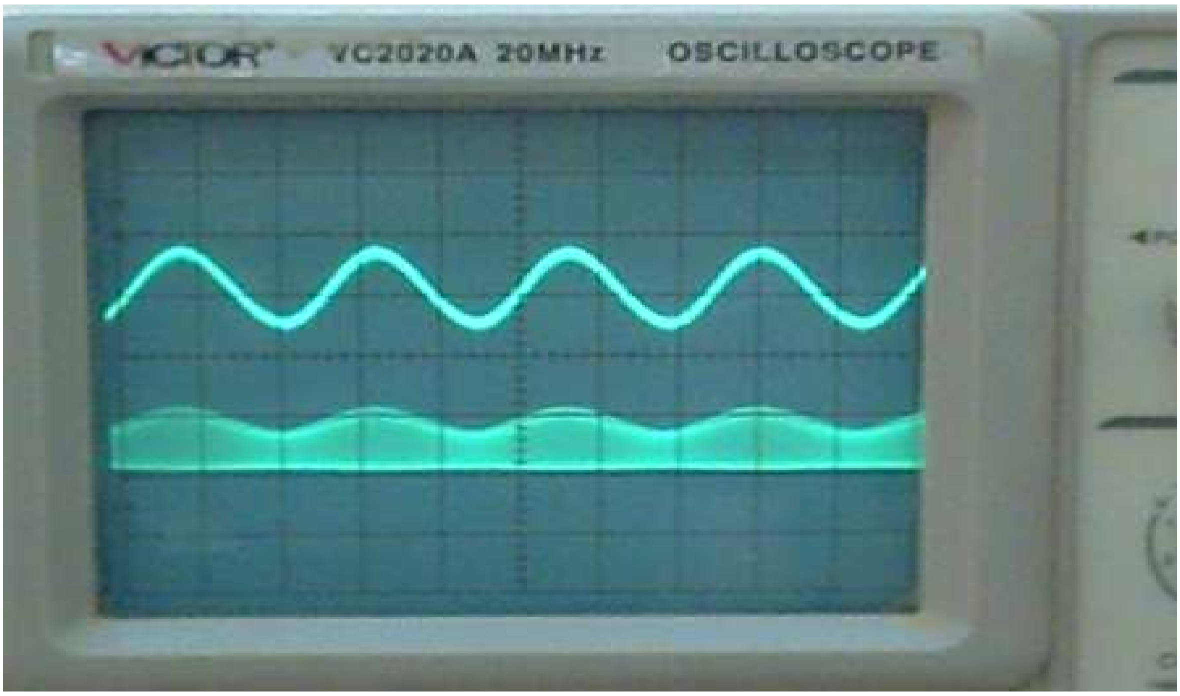


(b)

Figure 2 : The received interference signal
.a) without desired signal
b) with desired signal. 1volt/ Div. 1msec./ Div



(a)



(b)

Figure 3 : The H-signal, output of the multiplier .
a)with desired signal b) without desired signal. 1 Volt/Div. 1msec./Div.

References:

[1] Bernand W., et al, "Adaptive Noise Cancelling : Principles and Applications" Proceedings of the IEEE , Vol.63 , No. 12 , December 1975.

[2] Don Torrieri , " *principle of spread spectrum communication systems*" Springer , 2005.

[3] Peter Stavroulakis," *interference Analysis And Reduction For Wireless Systems*" Artech House , 2003

[4] Anton et al," *Calculus* ", John Wiley & sons, p78, 2005.

[5] Maurice J.Bouvier. "The Rejection of Large CW Interference in Spread Spectrum System". IEEE Trans. Commun. Vol. 26 , Feb., 1978.

[6] Viterbi A. J., *Principle of Coherent Communication*, McGraw-Hill , 1966.

[7] Ferrel G. Stremler, *Introduction to Communication Systems*, 3rd edition, Addison Wesley Longman, 1992.

[8] William F. Egan , *phase – lock Basics* , John Wiley & Sons , 1998

[9] Donland R. Stephens, *Phase-locked loop for wireless communication*, Kluwer academic publishers , 2002.