

## Analysis and implementation of carrier recovery process in a multimission remote sensing satellite ground station

*A separate data rate specific demodulator is required for each satellite-payload, if an earth station has to support data reception from multiple remote sensing satellites. A multimission demodulator would provide a low-cost solution to this problem.*

**By P.K. Jain, Surendra Pal and V.M. Pandharipande**

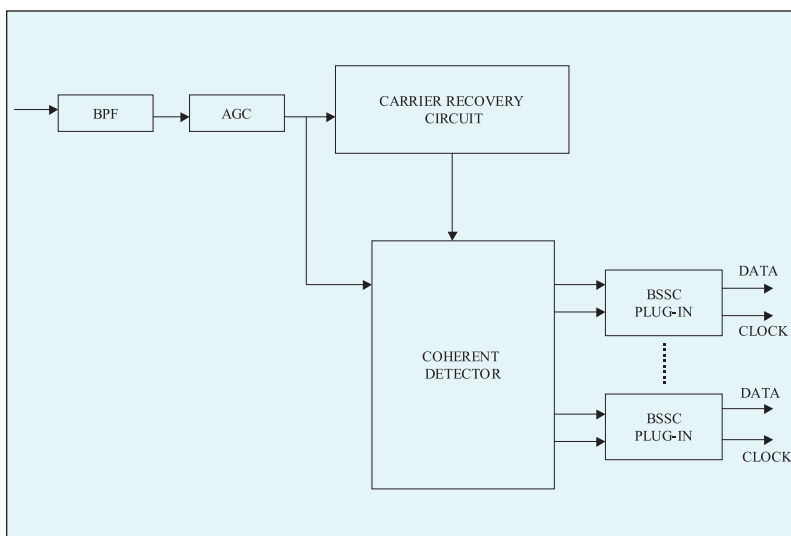


Figure 1: Configuration of a multimission demodulator

Remote sensing satellites transmit digitally modulated signals of different data rates in different PSK modulation schemes. The ground stations meant for receiving these signals are designed with the feed and front-end system broad enough to accommodate the reception of signals anywhere in the assigned spectrum.

A separate data rate specific demodulator has to be used for each satellite-payload, if mission specific design is used. This increases the system complexity and cost. A demodulator with multimission capabilities, therefore, provides a low cost solution to the problem of receiving multiple high data rate signals at different data rates in QPSK/UQPSK/BPSK schemes.

The design of the coherent carrier recovery circuit for the demodulation of suppressed carrier PSK signals is very critical and involves several performance considerations and trade offs. Various carrier regeneration schemes have been developed and implemented in this multimission demodulator design for optimal performance in different applications.

This article briefly describes these carrier recovery schemes, which are suitably configured to cater for multimission requirements of a remote sensing satellite ground station. The effect of the filtering and nonlinearity before carrier-referencing PLL used for the carrier recovery has also been analyzed in detail.

### Configuration of the multi-mission PSK demodulator

The simplified configuration of the multimission demodulator used for the remote sensing satellite data reception is shown in figure 1.

The demodulator consists of an input AGC amplifier and coherent carrier recovery and coherent detector circuits as its major constituents. The IF-signal-plus-noise is bandpass filtered, automatic gain control amplified and routed in parallel to the carrier recovery circuit and the coherent data detector.

The carrier recovery circuit regenerates the coherent reference for the demodulation and is routed to the data detector. The coherent data detector extracts the in-phase (I) and quadrature (Q) data streams, which are lowpass filtered and fed to the corresponding plug-in of the bit-synchronizer and signal conditioner (BSSC) unit.

The BSSC unit recovers the coherent symbol timing for synchronizing the data to the symbol timing clock. The outputs of this unit are the required serial data and bit rate clock.

While the demodulator part of the subsystem employs a common circuitry for different data rates of reception, different bit-synchronizer plug-in modules are required for different data rates.

### Coherent carrier regeneration analysis

The coherent carrier recovery from received modulated signal involves the modulation wipe-off through the nonlinear process, and recovering the

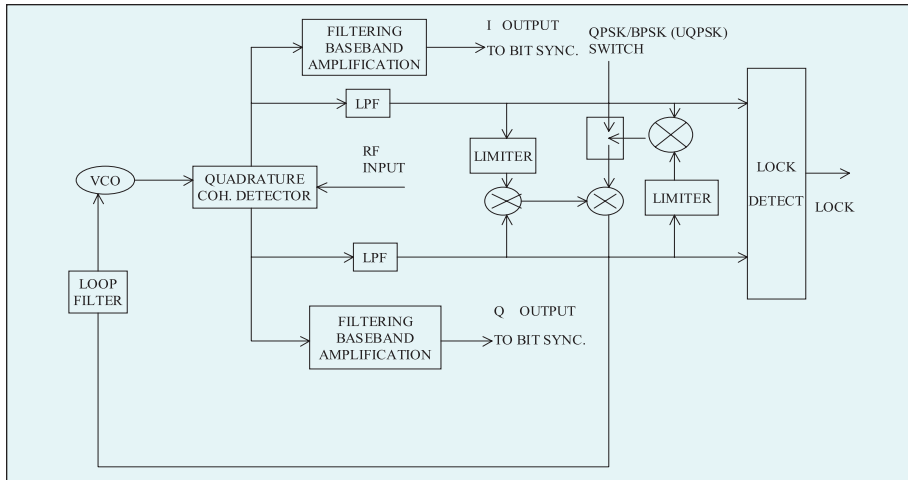


Figure 2: Costas loop carrier recovery for multimission demodulator

coherent carrier through a phase lock loop (PLL). The adaptability of this circuit for use in multimission data reception faces several design trade-offs in overall performance.

Before considering the design options for carrier regeneration, a review of the effects of nonlinearity on the signal-to-noise-ratio (SNR) of the regenerated coherent reference is provided. The discussion is mainly focused on BPSK carrier recovery for the simplicity though the analysis holds good for other PSK schemes as well. The analysis will be the basis on which the hardware implementation is done and the circuit is realized.

The received modulated signal is filtered and passed through an even-law nonlinearity (or times  $M$  multiplier) prior to phase referencing in a PLL.

The phase jitter and the static phase error on the regenerated carrier are the deciding factors of the detection efficiency of the demodulator and are affected by the filtering and the nonlinear effects of the carrier recovery process before the PLL. The fundamental expression relating the mean squared phase jitter in a PLL preceded by squaring (for BPSK demodulation) is given as

$$\sigma_{\phi}^2 = 1 / (\rho_L \cdot S_L) \quad (1)$$

Where  $\sigma_{\phi}^2$  is the mean squared output phase jitter of the recovered carrier,  $\rho_L$  is the SNR in the loop and  $S_L$  is the squaring loss.

It is evident from equation (1) that the squaring loss  $S_L$  controls the quality of the recovered carrier and requires thorough analysis to minimize its effect.

The filtering and nonlinear processing of the input modulated data waveform with noise results in three components of the signal-plus-noise entering into the PLL used for carrier phase referencing.

They can be listed as:

- The carrier component at  $2W_C$  (BPSK mode);
- The noise component  $N_{sq}$  representing  $S \times N$  and  $N \times N$  products; and
- The component representing the data self noise  $n_s(t)$ .

While the carrier component  $2W_C$ , after squaring, is the one to which the PLL has to lock on and track, the variance of the loop tracking error caused by squaring and self-noise will result in additional jitter in the recovered carrier and, consequently, the reduction in the SNR in the loop.

After the multiplication, while the excess noise contributes to the squaring loss by resulting in  $S \times N$  and  $N \times N$  cross products, the filtering effect on the data waveform also contributes to squaring loss through data self noise. Hence, the squaring loss can be related to the signal  $E_b/N_o$ , the filter noise bandwidth and shape. Lindsey and Simon have shown that for a particular filter type and bit waveform, an optimal filter bandwidth relative to the bit rate exists for each  $E_b/N_o$ . Putting together all the above factors, an expression for squaring loss can be derived as under

$$S_L = \{ (h_1/\alpha_s) + \{ (h_2 T_b/\alpha_s) / (2 E_b/N_o) \} \}^{-1} \quad (2)$$

Where,  $T_b$  is the bit duration and  $\alpha$  is  $B_{IF} / R_s$  where  $R_s$  is the symbol rate and  $B_{IF}$  is the double sided IF bandwidth.

$$h_1 = 1/2\pi \int_{-\infty}^{+\infty} S_m(\omega) |H(\omega)|^4 d\omega \quad (3)$$

$$h_2 = 1/2\pi \int_{-\infty}^{+\infty} |H(\omega)|^4 d\omega \quad (4)$$

where  $H(\omega)$  is the low pass equivalent of the input band pass filter transfer function and  $S_m(\omega)$  is the spectrum of the modulated signal.

It is seen from the above relation

that if the input pre filter to the nonlinearity is too wide relative to  $1/T_b$ ,  $h_1$  and  $h_2$  will increase, while if it is too narrow a decreases and both of these increase the squaring loss.

However, it can be deduced that after carrier regeneration, we need to consider only the power spectral density of the noise centered on  $2W_C$  at the output of the squarer. We can neglect  $S \times S$  product, such as the data self noise, considering the input filter bandwidth being much wider than the bit rate of the modulated signal for multimission operations. The output signal-to-noise-ratio  $(S/N)_{out}$  of the squarer is given by

$$(S/N)_{out} = (1/4) (B_{IF}/B_L) (C/N) / [1 + \{ (1/2) (C/N)^{-1} \}] \quad (5)$$

Where  $B_L$  is the noise bandwidth of PLL,  $C/N$  is the input  $C/N$  of the squarer, and  $B_{IF}$  is the noise bandwidth of the pre filter to nonlinearity.

Similarly, the equation of  $(S/N)_{out}$  for QPSK nonlinearity (quadrupler) is given by

$$(S/N)_{out} = (1/16) (B_{IF}/B_L) (C/N) / [1 + 4.5(C/N)^{-1} + 6(C/N)^{-2} + 1.5(C/N)^{-3}] \quad (6)$$

UQPSK modulation is different than QPSK in that the power in the I and Q channels is made unequal (usually in 1:5 ratio with Q channel takes lesser power than I channel) and, hence, the modulation is unbalanced to accommodate different data rates in I and Q channels. Now, if the total power in the spectrum is  $P$  and  $\rho$  is the fraction of  $P$  on I channel, then the output-signal-to-noise-ratio in the loop bandwidth of  $B_L$  of the UQPSK carrier referencing PLL can be derived as

$$(S/N)_{out} = \{ (1/4) (B_{IF}/B_L) (C/N) (2\rho - 1)^2 / [1 + (1/2) (C/N)^{-1} + \{ \rho(1-\rho) (1 - (R_Q/3R_I)) \} (C/N)] \} \quad (7)$$

Where,  $R_I$  is the data rate on I-channel and  $R_Q$  is the data rate on Q-channel. Equation (7) is the same as that for BPSK with squaring except for two additional terms.

The numerator has  $(2\rho-1)^2$  factor multiplying the input  $C/N$  ratio, which means that not all power is in I channel. The denominator has an extra term,  $\{ \rho(1-\rho) (1 - (R_Q/3R_I)) \} (C/N)$ , which is the result of  $S \times S$  cross talk introduced by the UQPSK carrier regeneration process.

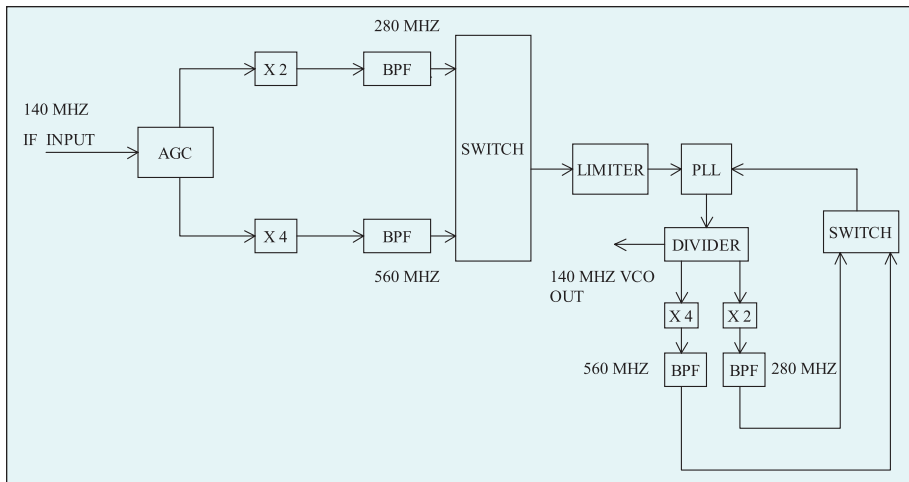


Figure 3: Multiplication carrier recovery for multimission demodulator

The desired  $2W_c$  term to be tracked by the PLL is proportional to the total signal power entering the nonlinearity, and the  $S \times S$  plus noise is also proportional to the signal power in the spectrum. With the increase in the input SNR to the PLL, the loop gain increases and, consequently, expands the noise bandwidth. Similarly, the self-noise spectral density due to the  $S \times S$  product also increases. Thus, the output noise variance of the loop also increases.

However, incorporating a limiter in front of the PLL will, to some extent, provide a solution to the problem of varying input SNR and resultant jitter in the PLL at high input SNR. The inclusion of a limiter introduces a signal suppression factor into the analysis of the loop that improves the squaring loss performance of the loop at higher  $E_b/N_0$  ratios.

The limiter in front of the loop not only keeps the loop parameters constant, but also keeps the SNR of the signal entering into the loop constant under high  $E_b/N_0$  ratio conditions. The

effect is the constant variance of output phase jitter.

But the case is different at very low SNR and the limiter degrades the SNR performance of the signal.

As already discussed, the jitter and static phase error in the recovered carrier depends on its SNR and results in the detection loss of the demodulator. A relation can be established between the detection loss, variance of the phase noise of the reconstructed coherent reference and the probability of bit error in the demodulated data, as under

$$L \text{ (dB)} = (4.34/\rho_L) \{ (1+2\gamma) \{ 1 + \{ (1+2\gamma) / (2\rho_L) \} \} \} \quad (8)$$

Where,  $\rho_L$  is the SNR in the loop,  $\gamma = Z_2 * E_b/N_0$ , and  $Z_2$  is the effective loss in  $E_b/N_0$  caused by synchronization error.

Plots relating these parameters are available, serving as the guidelines to determine the required SNR of the referenced carrier to provide the BER close to theory for the chosen detection loss.

For UQPSK, the fraction of power  $\rho$

in I channel will become the guiding factor to determine the requirement of SNR of the referenced carrier, for a given detection loss. The fraction power  $\rho$  decides the cross-talk between I and Q channel in that as  $\rho$  gets closer to one, I channel derives more power from the total spectral power and the Q channel gets severely affected by the cross-talk from I channel.

Similarly, as  $\rho$  deviates more and more from unity, cross talk from Q channel starts affecting the I-channel. One can use above referred QPSK plots as the starting point and add in extra margin as a function of  $\rho$  and derive the required SNR of carrier reference for corresponding BER performance, for the given loss.

### Various techniques of implementation

The coherent carrier recovery circuit's — being the most responsible constituent of the demodulator — design for multimission features has to be done by keeping various critical requirements in view. The effects of input data rate and SNR variations on the overall performance of the system have to be considered, and system elements have to be designed with broadband characteristics.

The following schemes are considered for realizing the coherent carrier recovery process in a multimission demodulator, as they are effectively configurable for varying requirements:

- Fourth-order Costas loop with a selection for second order;
- Multiplication loop with a selection between fourth power and squaring devices; and
- Long-loop PLL carrier recovery as an extension of the multiplication loop.

### Costas loop

An explanation of this scheme is done with reference to the block diagram in figure 2.

The input modulated signal (received from the automatic gain controller (AGC) amplifier) is split into two paths and are mixed with the quadrature local oscillator signal (the VCO of the PLL) in the quadrature coherent detector to produce the desired I and Q base-band signals.

These signals are amplified in broadband DC coupled amplifiers and given to the bit-synchronizer for further processing. The samples of the same I and Q data streams are also amplified and

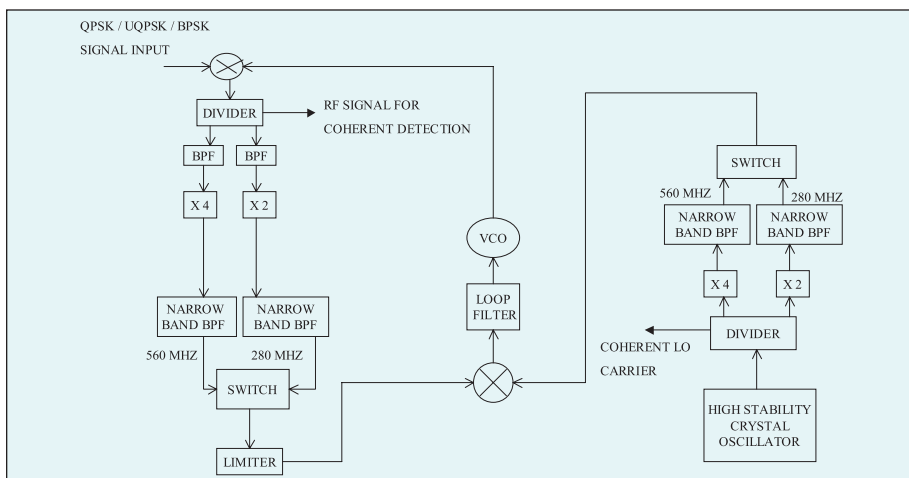


Figure 4: Long-loop configuration for carrier recovery

filtered in the Costas-arm-filters. The arm-filter outputs are fed to the Costas phase detectors, which generate a phase error estimate through a hard limited polarity loop.

The loop has two operating modes, one as a carrier reference loop for BPSK (or UQPSK) and the other is for QPSK carrier regeneration. Effectively, when BPSK

mode is selected, the loop realizes the second order nonlinearity. When QPSK mode is selected, it realizes the fourth order nonlinearity. The corresponding output phase error estimate is fed to a PLL integrator or loop-filter. The integrator output drives the VCO. The VCO output through the quadrature hybrid becomes the local oscillator (LO) inputs to

the mixers of the coherent detector.

While the selection of the order of the nonlinearity facilitates the demodulator to function for the reception of data from BPSK/UQPSK/QPSK modulations, the performance of the demodulator for receiving signals with different data rates and  $E_b/N_0$  ratios mainly depends upon the bandwidth and characteristics of the Costas arm filters.

As has already been discussed, at higher input data rates and filter noise bandwidth approximating the symbol rate, the  $S \times S$  data self noise increases and, though the SNR is controlled by the hard limiter, contributes to the increase in loop jitter.

This results in the requirement of higher SNR in the regenerated carrier for meeting the BER performance close to the theory. At lower symbol rates ( $B \gg 1/T_s$ ), the excess noise bandwidth of the arm filters contributes to increased squaring loss due to  $S \times N$  and  $N \times N$  products increasing.

To make the loop perform with reasonable phase jitter in the recovered carrier reference over the range of input symbol rates, the arm filters should be chosen with a noise bandwidth optimized for mid-range symbol rates such that the compromised performance is achieved between data self noise and squaring loss, and the performance degradation due to phase jitter may not be more than 2 dB at the extremes of the symbol rate range.

### Multiplication loop

Multiplication loop with switching between the second order for UQPSK/BPSK carrier recovery and fourth order for QPSK carrier recovery is given in figure 3.

The analysis is identical to the analysis of Costas loop as explained above. The noise performance of both the loops is identical.

The one-sided lowpass arm filters are replaced by the double sided band-pass filter centered at the ( $\times M$ ) IF carrier frequency. The double sided noise bandwidth of this pre filter is chosen for a mid-range symbol rate. The noise bandwidth of the pre filter for second order loop is chosen for the I-channel data rate of UQPSK signal. Selection is done between second order and fourth order output into the phase lock loop carrier referencing. The VCO output is also simultaneously selected either for  $x2$  or  $x4$  multiplied output.

Sl. No	C/No at Demod. input (dBHz)	Data Rate 20.8 MBPS (QPSK)			Data Rate 105 MBPS (QPSK)		
		$E_b/N_0$ at Demod. input (dB)	BER with conventional carrier recovery	BER with Long Loop Carrier Recovery	$E_b/N_0$ at Demod. input (dB)	BER with conventional carrier recovery	BER with Long Loop Carrier Recovery
1	81	7.8	-	$5 \times 10^{-3}$	-	-	-
2	82	8.8	-	$2 \times 10^{-3}$	-	-	-
3	83	9.8	-	$3 \times 10^{-4}$	-	-	-
4	84	10.8	$6 \times 10^{-5}$	$4 \times 10^{-5}$	3.8	-	$4 \times 10^{-2}$
5	85	11.8	$8.5 \times 10^{-6}$	$8 \times 10^{-6}$	4.8	$4 \times 10^{-2}$	$2.5 \times 10^{-2}$
6	86	12.8	$1.5 \times 10^{-6}$	$1.1 \times 10^{-6}$	5.8	$2 \times 10^{-2}$	$1.5 \times 10^{-2}$
7	87	-	-	-	6.8	$7.5 \times 10^{-3}$	$7 \times 10^{-3}$
8	88	-	-	-	7.8	$3 \times 10^{-3}$	$2.5 \times 10^{-3}$
9	89	-	-	-	8.8	$1.2 \times 10^{-3}$	$1 \times 10^{-3}$
10	90	-	-	-	9.8	$3 \times 10^{-4}$	$2.5 \times 10^{-4}$
11	91	-	-	-	10.8	$5.1 \times 10^{-5}$	$5 \times 10^{-5}$
12	92	-	-	-	11.8	$6.8 \times 10^{-6}$	$7 \times 10^{-6}$
13	93	-	-	-	12.8	$1 \times 10^{-6}$	$9 \times 10^{-7}$

**Table 1: Comparative BER performance of conventional and long loop PLL QPSK carrier recovery techniques**

The VCO output is power divided and one part is taken as the coherent carrier reference.

### Long-loop PLL carrier recovery

As discussed above, the wideband pre-filtering prior to the nonlinearity for modulation wipe-off has to be done to accommodate the highest data rate modulated signals for minimum squaring loss. The excess noise bandwidth in the filter, while receiving the lower data rate modulated signals, increases the squaring loss and, consequently, reduces the pre-PLL SNR.

Therefore, the acquisition threshold of the carrier referencing PLL vis-à-vis the input ratio increases, resulting in reduction in the link margin while receiving the lower data rate modulated signals. A relatively new and unique technique, which uses the multiplication loop, in long-loop configuration, to recover the coherent carrier, has been developed.

The method, while facilitating the use of wider pre-filter at the input carrier recovery nonlinearity, also optimizes the SNR at the input of the carrier referencing PLL by employing a narrow bandpass filter. This scheme has provided better lock threshold performance and loop stability in regenerating the coherent carrier for low SNR at different modulation schemes and input data rates, without degrading the BER

Sl. No	C/No at Demodulator input (dBHz)	$E_b/N_0$ at Demodulator input (dB)	BER with conventional carrier recovery	BER with Long Loop Carrier Recovery
1	84	5.2	-	$3 \times 10^{-2}$
2	85	6.2	$1.8 \times 10^{-2}$	$1.5 \times 10^{-2}$
3	86	7.2	$8.6 \times 10^{-3}$	$6.9 \times 10^{-3}$
4	87	8.2	$2.8 \times 10^{-3}$	$2.3 \times 10^{-3}$
5	88	9.2	$6.1 \times 10^{-4}$	$6 \times 10^{-4}$
6	89	10.2	$8.7 \times 10^{-5}$	$8.5 \times 10^{-5}$
7	90	11.2	$1.5 \times 10^{-5}$	$1.4 \times 10^{-5}$
8	91	12.2	$2.8 \times 10^{-6}$	$2.5 \times 10^{-6}$
9	92	13.2	$4.2 \times 10^{-7}$	$3 \times 10^{-7}$

NOTE: A power split-loss of 7 dB in the Q-channel of UQPSK modulated Signal (due to 20% of HPA power in Q-channel) has been included while calculating the effective  $E_b/N_0$  at the Demodulator input.

**Table 2: BER performance of conventional and long loop PLL BPSK/UQPSK carrier recovery techniques for the 15.062 Mbps data rate (UQPSK, Q-channel)**

performance of the demodulator.

The configuration of the long-loop scheme is explained with reference to figure 4.

The received modulated RF signal is mixed with the PLL VCO signal to provide an IF signal. The signal is passed through a three-way power divider. One output from the divider

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goes for the coherent detection, while the other two outputs are fed to the corresponding bandpass filters with the bandwidth selected optimally for minimum squaring loss for the highest baud rate, in each of QPSK/UQPSK (BPSK).

The filtered signal is then passed through the nonlinearity to remove the modulation. The fourth or second harmonic component of IF carrier frequency, after the nonlinearity, is filtered through the corresponding very narrow bandpass filter.

The output from the narrow bandpass filter is then phase locked with the  $\times 4$  or  $\times 2$  multiplied component of a very highly stable crystal oscillator source. The crystal oscillator source frequency is centered at the IF carrier frequency and becomes the coherent carrier reference for data detection after phase locking.

The PLL tracks the Doppler frequency variations of the input signal by correcting the VCO output frequency accordingly, while maintaining the IF output of the mixer at a constant frequency.

The VCO is swept for initial signal acquisition and is designed for acquiring the signal with a Doppler shift as high as  $\pm 500$  kHz. As the Doppler effect is compensated prior to the carrier recovery non linearity, the bandwidth is controlled by using a narrow bandpass filter at the input of the carrier referencing PLL. This becomes possible in this configuration and the SNR at the output of the nonlinearity can be maximized to improve the overall lock acquisition threshold of the PLL.

### Comparative BER performances

Table 1 and table 2 compare the BER performances of the conventional and long-loop carrier recovery methods, practically measured with the identical input pre-filters to the nonlinearity, and at different data rates in different modulation modes, in a remote sensing satellite ground station.

The noise bandwidths of the input bandpass filter prior to the QPSK and BPSK nonlinearity are 116.14 MHz and 222.03 MHz, respectively. Similarly, the noise bandwidths of the pre-PLL narrowband filter and PLL are 5.04 MHz and 100 KHz, respectively.

The noise-bandwidth of the input filter prior to BPSK nonlinearity is chosen such that the SNR profile at the output of the squarer at a specific data rate remains almost same as that of

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the QPSK configuration at the same symbol rate and, therefore, maintains the same SNR at the input of the PLL.

As seen in the table 1, the entries pertaining to the BER with conventional carrier recovery for the data rate of 20.8 Mbps starts from the  $E_b/N_0$  of 10.8 dB against the entries of long-loop carrier recovery, which begins with the  $E_b/N_0$  of 7.8 dB itself. This is due to the poor acquisition threshold of approximately 10.5 dB of the conventional loop as against the 7.5 dB of the long-loop PLL, which starts recovering carrier from the lower  $E_b/N_0$  while allowing the noise bandwidth of the pre-filter to the nonlinearity to be wide enough (116.14 MHz) to accommodate the data rate of 105 Mbps. The case with other entries in table 1 and table 2 at other data rates and modulation schemes is similar.

### Conclusion

Various design options are considered to realize the carrier recovery process in a multimission demodulator. The Costas loop and the multiplication loop configurations, among these, are the conventional techniques.

The analysis and the results of practically designed unit using these conventional design procedures show that the unit performs reasonably well for the limited range of input data rate (such as 40 Mbps to 110 Mbps) and  $E_b/N_0$  variations. For the wider input parameter variations, these techniques require a higher acquisition threshold, and result in signal loss at lower antenna elevation angles below 5 degrees.

Very large Doppler offset and rate in the received signal from remote sensing satellites impose more restrictions on the PLL design. The long-loop technique has given solutions to these design restrictions while providing overall improvement in the carrier recovery loop. The

demodulator with this technique acquires lock, and starts regenerating the coherent carrier at very low input SNR, a feature very useful in remote sensing satellite ground station applications where the coherent continuous clock is required to be supplied to the recording system even from the antenna angles lower than one degree in elevation.

The BER performances of all the carrier recovery methods, however, are found to be almost identical.

**RF**

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