Digital Auto Frequency Control for a General Purpose IF Subsystem with Multi-Modulation Schemes

X. Chen and H. Zhang

***Abstract*—This paper presents a general purpose auto-frequency control (AFC) algorithm for wireless communication systems with multi data rate and multi modulation schemes, such as FSK, GFSK, and MSK, etc. The proposed algorithm is a non-decision-aided recursive closed-loop which does not require timing recovery and permits wider tracking range. By implementing an adaptive loop gain control, the proposed AFC can achieve a higher tracking speed as well as better performance. A novel normalization scheme based on digital RSSI and bit shifting /truncation is also included to remove the effects on the estimated carrier frequency offset (CFO) by the gains and distortions all along the receiving path, such as inter-channel interference (ICI) and Auto-Gain Control (AGC) uncertainties which are not taken into consideration in most of the existing references.**

***Index Terms*—Adaptive AFC, CFO, RSSI, synchronization, normalization.**

I. Introduction

Many digital frequency and phase modulation systems through wireless channels such as frequency shift keying (FSK), Gaussian frequency shift keying (GFSK), minimum phase shift keying (MSK) etc. are sensitive to carrier frequency offset (CFO) caused by transceiver oscillator instability and/or Doppler shift especially when data is transmitted in a burst mode [1]. One possible solution to solve this problem is applying an auto-frequency calibration (AFC) block in the receiver to automatically estimate and compensate for such frequency offset. However, comparing with plenty of design and research on FSK/GFSK/MSK transceivers that can be found in literatures [2]–[5], there are very limited articles on studying CFO estimation and compensation for such systems.

Some studies on CFO estimation and/or compensation can be found in literatures [6]–[12]. F. Natali summarized a set of analog AFC tracking algorithms in [6], which can be recognized as the basis of the modern digital AFC. A set of digital closed-loop decision-aided AFC tracking algorithms for GFSK systems are proposed in [8] and [9], where [9] is a modified version of [8]. Both of these two algorithms require reconstruction of transmitted data symbols submitting to the CFO estimator as reference information. Therefore the trackable AFC range is limited no to exceed the maximum frequency divination, and accurate sample timing recovery is required. In [5], an open-loop AFC tracking algorithm which directly estimates the DC-offset of the discriminator output is proposed. However, this AFC algorithm is only applicable for frequency modulation system with discriminator demodulator. A direct CFO estimator based on received signals and re-modulated transmitted symbols is proposed. But the channel response and the training sequence must be known in advance. Meanwhile, some AFC algorithms such as FFT based and Maximum Likelihood can also be found in literatures [11], [12].

On the other hand, most of the existing AFC algorithms assume that the received signal has constant envelope. How-ever this assumption is not always true if the Inter-channel Interference (ICI) and AGC uncertainties are taken into ac-count. Luo and Ma invented a normalization method for AFC in GFSK systems in [13]. The estimated CFO is normalized to the maximum deviation and hence it becomes irrelative to the gains along the receiving path.

In this paper, a novel non-decision-aided digital closed-loop AFC tracking algorithm based on the moving average of the digital discriminator's output is proposed. This AFC does not require timing recovery and /or source data recovery. The existing decision-aided AFC methods usually require the transmitted data being demodulated even before CFO is compensated, thus the tracking range is limited to below the maximum frequency deviation. The proposed AFC does not have such requirement and thus has wider tracking range. An adaptive tracking loop gain scheme is also proposed to achieve a faster and accurate tracking. It automatically switches between Fast and Accurate modes by adjusting the loop gain according to the estimated CFO for each iteration. Lastly, the ICI and AGC uncertainties are also taken into consideration by normalizing the I Q samples by re-using the existing digital RSSI. Meanwhile, the normalization is based on a simple bit shift and truncation process which makes it easy to implement.

The remaining parts of this paper are organized as follows. In Section II the mathematical system model for the general continues phase shift key systems is presented. Thereafter, the proposed AFC algorithm is presented in Section III. The details of three main novel components are discussed in the three subsections of this section. The comparison to the state of the art and discussions are presented in Section IV, and the simulation results are demonstrated in Section V. Finally the conclusion is drawn in the last section.

II. System Model

To simplify the discussion, a common continues-phase frequency shift keying (CPFSK) modulation scheme is considered in this section. However, the proposed AFC algorithm is applicable to frequency modulation scheme including FSK, GFSK, and MSK. It will be shown that the presented mathematical derivations based on general CPFSK

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system can be easily applied to the other phase modulation schemes mentioned above.

Assume the received CPFSK signal is not distorted by channel and pre-detection filter, it can be denoted as

$$x(t) = \sqrt{2E_b / T_b}\cos(2\pi f_c t + \theta(t) + \theta_0) + n(t) \qquad (1)$$

where f_c is the carrier frequency, $n(t)$ is an additive bandpass Gaussian noise with one-sided power spectrum density N_0, E_b and T_b are the bit energy and bit period respectively. It should be noted that E_b is the bit energy with the effects of AGC and with ICI removed by low-pass filter. It varies with different AGC or different ICI. θ_0 is the initial phase offset and $\theta(t)$ is the frequency modulated phase as shown below

$$\theta(t) = 2\pi h \int_{-\infty}^{t} \sum_{n=-\infty}^{\infty} x[n] g(\tau - nT) d\tau \qquad (2)$$

where $g(t)$ is the pulse shaping function for binary data, and h is the modulation index.

III. The Proposed AFC Algorithm

In this section, a closed-loop recursive automatic frequency calibration algorithm for CFO estimation and compensation is presented. The following Fig. 1 demonstrates the block diagram of the proposed AFC.

The received waveform is first down converted to produce baseband in-phase (I) and quadrature (Q) signals. Ignoring the noise, they can be expressed as

$$I(t) = \sqrt{2E_b / T_b}\cos(2\pi\Delta ft + \theta(t) + \theta_0) \qquad (3a)$$

$$Q(t) = \sqrt{2E_b / T_b}\sin(2\pi\Delta ft + \theta(t) + \theta_0) \qquad (3b)$$

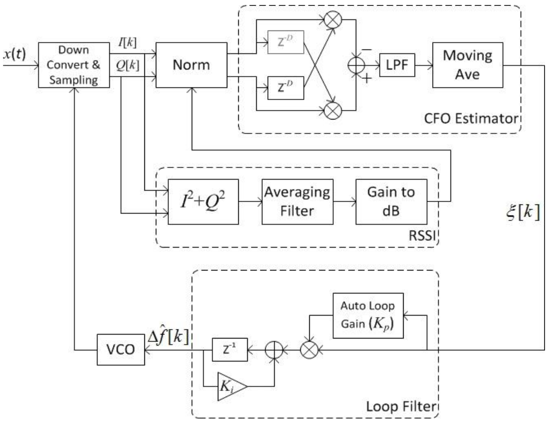

Fig. 1. Block Diagram of the Proposed AFC.

where $\Delta f = f_c - f'_c$ is the frequency offset, and f'_c is the frequency generated by the local oscillator.

These $I(t)$ and $Q(t)$ signals are then sampled at sampling rate $1/T_s$ and normalized by the normalization block. The normalized samples are then passed to the digital discriminator whose delay taps is set to be D. Thereafter a moving average block with window size L_w generates an indication of the difference between the transceiver's carrier frequency offset Δf. Lastly, the error signal is filtered to smooth out the noise and used to steer the PLL to generates the local oscillator's frequency towards f_c. The details of each function block are discussed as below.

A. Normalization

From Fig. 1, it can be imagined that the error signal is proportional to the gains along the receiving path including AGC adjustment and low-pass filter for removing ICI. It is necessary to remove or suppress such correlation by normalization. In [6] a normalization scheme based on peak detection is proposed. In this section a novel normalization scheme by reusing the digital RSSI is proposed.

The over-sampled version of the complex baseband signals given by Eqn.(3a) and (3b) can be written as.

$$I[k] = I(kT_s) = \sqrt{\tfrac{2E_b}{T_b}}\cos(2\pi\Delta fkT_s + \theta(kT_s) + \theta_0) \quad (4a)$$

$$Q[k] = Q(kT_s) = \sqrt{\tfrac{2E_b}{T_b}}\sin(2\pi\Delta fkT_s + \theta(kT_s) + \theta_0) \quad (4b)$$

These I Q samples are to be passed to the discriminator for further operation. However, the discriminator's output is proportional to the overall gains along the receiving path. Thus the estimated CFO in such way has a constant ambiguity which needs to be removed. Other than the normalization scheme of [6], the proposed normalization method directly normalizes the I Q samples before them being fed to the discriminator. This is to narrow down the required dynamic range of the digital discriminator and achieve a stable tracking speed. Meanwhile the reusing of RSSI helps to save power consumptions and chip areas.

In a digital RSSI, the signal power is estimated by the average value of the powers of I Q samples, i.e.,

$$\widetilde{P} = \sum_{k=0}^{L_R-1} I^2[k] + Q^2[k] \qquad (5)$$

where LR is the length of the observation window of the RSSI. The estimated signal power is mapped to dB with resolution of 1 dB and fed to the Normalization block to normalize the I Q samples. Since the dividing operation is too complex to realize, the proposed normalization is based on truncating and shifting the fixed point samples. Each integer value of RSSI corresponds to a number of bits and direction of I Q samples (fixed-point numbers) are to be shifted. Then the shifted numbers are truncated according to the system requirement. This shift and truncation process should be easily implemented, and the I Q samples are normalized to certain range so than the AFC will never lose convergence given other necessary conditions are satisfied.

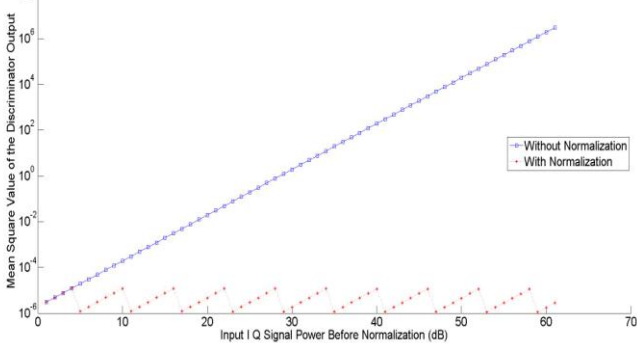

Fig. 2. Discriminator output of the normalized I Q samples.

Fig. 2 illustrates the output of the digital discriminator where the input I Q samples are normalized by the proposed normalization scheme. The power of the input signal spans from 0 to 60 dB (normalized to V_{pp} = 1V). After normalization, it is clear that the discriminator's output is limited in a reasonable range.

B. Estimation of Carrier Frequency Offset

Consider the digital discriminator's output whose inputs are the normalized I Q samples and substitute to Eqn.(2), it can be expressed as below.

$$\begin{aligned}\xi[k] &= I[k-D]Q[k]-I[k]Q[k-D] \\ &= \sin\{(2\pi\Delta fDT_s)+[\theta(kT_s)-\theta(kT_s-DT_s)]\} \\ &= \sin\{2\pi[\Delta fDT_s+h\phi(kT_s)]\}\end{aligned} \tag{6}$$

where φ (kT_s) is defined by

$$\phi[kT_s]=\int_{(k-D)T_s}^{kT_s}\sum_{n=-\infty}^{\infty}x[n]g(\tau-nT)d\tau \tag{7}$$

Consider the average of the above discriminator output ξ [k] with observation window size being L_w samples. If the preamble satisfies the condition of (0, 1) balance in the observation window, then the average output can be expressed as

$$\begin{aligned}\bar{\xi} &= \tfrac{1}{L_w}\sum_{k=l}^{l+L_w-1} I[k-D]Q[k]-I[k]Q[k-D] \\ &= \tfrac{1}{L_w}\sum_{k=l}^{l+L_w-1}\sin\{2\pi[\Delta fDT_s+h\phi(kT_s)]\} \\ &= \tfrac{1}{L_w}\left\{\begin{array}{l}\sin\{2\pi[\Delta fDT_s+h\phi(kT_s)]\}\sum\limits_{k=l}^{l+L_w-1}\cos[2\pi h\phi(kT_s)] \\ +\cos\{2\pi[\Delta fDT_s+h\phi(kT_s)]\}\sum\limits_{k=l}^{l+L_w-1}\sin[2\pi h\phi(kT_s)]\end{array}\right\}\end{aligned} \tag{8}$$

Under the condition of preamble being (0,1) balance, it can be proven that the first summation term of Eqn.(8) in the big bracket can be approximated by a positive constant, and the second summation term approximately equals to zero, i.e.,

$$\sum_{k=l}^{l+L_w-1}\cos[2\pi h\phi(kT_s)]\approx\alpha \tag{9a}$$

$$\sum_{k=l}^{l+L_w-1}\sin[2\pi h\phi(kT_s)]\approx 0 \tag{9b}$$

Therefore according to all of above and take the AWGN noise into consideration, the moving average in Eqn.(8) can be re-written as

$$\begin{aligned}\bar{\xi} &\approx (\alpha/L_w)\sin(2\pi\Delta fDT_s)+w(n) \\ &= 2\pi\beta\Delta fDT_s+w'(n)\end{aligned} \tag{10}$$

where $w(n)$ refers to the effect of AWGN noise, and $w'(n)$ is the overall noise including the approximation error. Thus, the frequency offset can be estimated by

$$\Delta\hat{f}\approx\frac{1}{2\pi\beta\Delta fDT_s}\bar{\xi} \tag{11}$$

The approximation in Eqn.(10) holds only when ($2\pi\Delta fDT_s$) is small. But if a feed-back tracking loop is employed, the condition for the loop to converge is that sin ($2\pi\Delta fDT_s$) has the same sign as Δf to prevent the estimation of CFO being tracked to a wrong direction. Therefore the condition of convergence is $|2\pi\Delta fDT_s|<\pi$, i.e.,

$$-\frac{1}{2DT_s}<\Delta f<\frac{1}{2DT_s} \tag{12}$$

C. Tracking Loop

In this subsection, the tracking loop with adaptive gain is presented. The structure of the proposed tracking loop is based on the standard feed-back loop except that the loop gain in the loop filter is adaptive to achieve fast tracking speed as well as accuracy.

Since the parameter β in Eqn.(11) would be affected by factors such as sample timing error, a closed-loop recursive method is implemented to avoid this problem. The overall gain $1/(2\pi\beta T_s)$ is absorbed into the loop gain K_p. It should be noted that K_p has a certain range of tolerance for tracking convergence. Hence the tracking speed and accuracy is not very sensitive to the error on β .

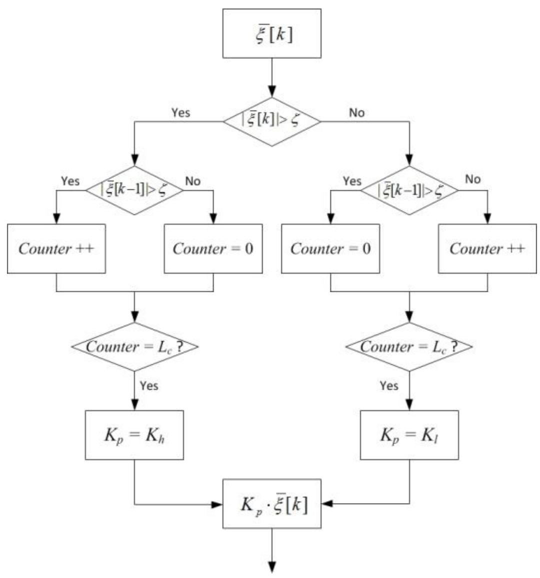

Fig. 3. Flow chart of automatic gain control.

The loop filter has two working modes, i.e., Fast Mode and Accurate Mode. It automatically switches between these two modes according to the absolute value of the moving average for each iteration, i.e., $\bar{\xi}[k]$ as shown in Fig. 1. A positive threshold ζ is preset. The tracking loop is switched to Fast or Accurate mode if $|\bar{\xi}[k]|$ is detected above or below ζ respectively. The mode switching is achieved by selecting High or Low values of the loop gain K_p. It is set to high value K_h in the Fast mode and low value K_l in the Accurate mode. K_h and K_l are preset for each system. The optimization of K_h and K_l is not discussed in this paper and could be further studied in future work.

In noisy cases, it is possible that $\bar{\xi}[k]$ jumps between above and below ζ respectively for some consecutive iterations. To avoid such gain oscillation, the mode switching can be limited to happen only when $\bar{\xi}[k]$ is stabilized after switching from one side of the threshold to another. Fig. 3 illustrates the flow chart the logics of gain control.

IV. Discussion

In this section, the main features and the novelties of the proposed AFC is discussed. The following Table I shows the comparison of the main features of the proposed AFC and existing algorithms.

From the contents listed in the table, it can be concluded that the proposed AFC algorithm has attractive features in the areas of tracking speed, accuracy, and trackable range. The main novelties of the proposed AFC are listed below:

1) The normalization scheme by reusing the existing digital RSSI and shifting / truncation based process to fixed point samples. This scheme keeps the balance between the performance and system complexity. The I Q samples are normalized to certain range by simple MSB searching, bit shifting and truncation.
2) The non-decision-aided CFO estimation algorithm which achiever wider trackable range. The CFO is estimated by the moving average of the discriminator outputs. It does not require timing recovery and does not need to reconstruct transmitted symbols as well. Therefore, there is no limitation on tracking range being less that the maximum frequency deviation. The trackable CFO range is identified mathematically in Eqn.(12). The contents listed in Tab. I also proves this attractive feature.
3) The adaptive loop gain. By automatically switching the loop gain between High and Low values according to the estimated CFO for each iteration, the loop gain is switched between Fast and Accurate modes accordingly. Therefore, the proposed AFC can achieve a higher tracking speed and better performance as well.

TABLE I: Comparison of Main Features

Parameters	Proposed	[10]	[9]	[6]	[4]
IF	172.8kHz	8MHz	8MHz	N.A.	910MHz
Data Rate	1.2-4.8kbps	3Mbps	3Mbps	0.7-2.1Mbps	1Mbps
Over Sampling	9X	8X	8X	N.A.	4X
Converge speed	16 bits	> 32 bits	> 32 bits	N.A.	N.A.
Modulation	Multi	GFSK	GFSK	GFSK	GFSK
Dec-Aid	No	Yes	Yes	N.A.	No
Norm	Yes	No	No	Yes	No
Adaptive Gain	Yes	No	No	No	No
Trackable $\Delta f/R_b$	±4.5	±0.033	±0.033	N.A.	±0.24
Residual Δf (SNR=10dB)	2.4%	3.0%	3.5%	N.A.	3.9%

V. Simulation Results

In this section, the Matlab simulations of the proposed AFC are provided. In our simulation, a GFSK system is simulated. The GFSK parameter for BT is 0.5, modulation index h is 1, and the filter length to be calculated is ±2 symbols span. The signal is transmitted through an AWGN channel. The RF and IF frequencies are 420 MHz and 172.8 kHz respectively. At the receiver side, ADC is used to convert the received analog waveform to digital samples. The sampling frequency for the ADC is 5.5296 MHz with oversampling ratio of 128. Thereafter, a set of CIC and low-pass filters are used which results in the final sampling rate of I Q signals are 43.2 kHz and thus the oversampling ratio is 9. The carrier frequency offset is 3.67 kHz, which is equivalent to ±4 ppm. The delay for the digital discriminator is set to 1.

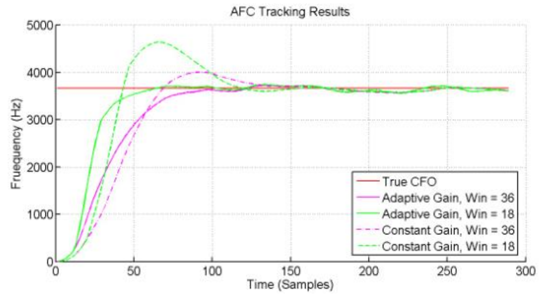

Fig. 4. Tracking logs of the AFC with different settings.

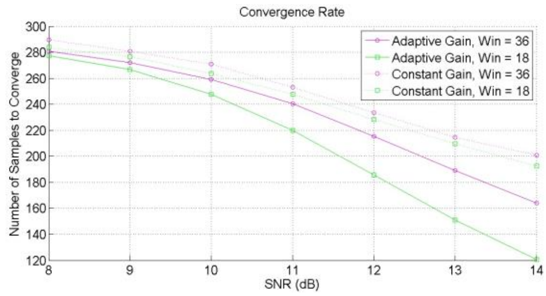

Fig. 5. Comparison of convergence speed with different settings.

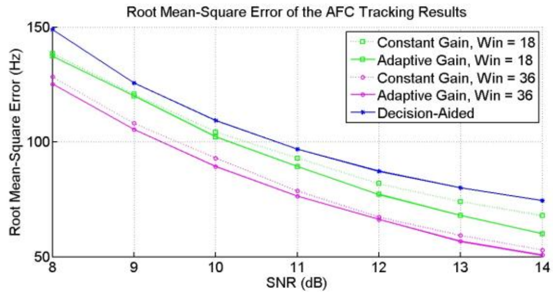

Fig. 6. Comparison of convergence speed with different settings.

Fig. 4 illustrates the tracking log of the AFC with and with-out adaptive gain control. Two different observation window lengths are considered. It can be observed that the convergence speed of the system with adaptive gain control is faster than that with constant loop gain. Meanwhile, this figure also shows the effects of the observation window length on the tracking speed. On one hand, the larger the window size, the higher the estimation accuracy. On the other hand, the larger window size also slows down the tracking speed. Therefore it has to be balanced.

Fig. 5 further demonstrates the convergence speed of the AFC with different setting.

Lastly, in Fig. 6 the tracking performance of the proposed AFC with different settings are presented. The performance of the decision-aided AFC in [9] is also simulated for comparison. It is shown that the proposed algorithm achieves better performance than then decision-aided algorithm. This is because the performance of the decision-aided AFC depends on the accuracy of timing recovery. Given the fact that there exists some random timing error, decision-aided AFC suffers some performance degradation.

VI. Conclusion and Future Work

A novel moving average based AFC tracking algorithm is presented. This algorithm can be applied to any CPFSK systems. The carrier frequency offset is estimated by averaging the digital discriminator's output. An adaptive tracking loop with auto-switching loop gain is proposed to achieve higher tracking speed and accuracy. A simple normalization scheme with the assist of the existing digital RSSI by shifting and truncation is also proposed to narrow down the required dynamic range the digital discriminator and removes the effects of inter-channel interference and AGC uncertainties. The simulation shown the proposed AFC has attractive features comparing with existing references.

On the other hand, the automatic loop gain control scheme proposed in this paper is a simple comparing and switching process. It is valuable to further optimize the threshold and the values of High and Low gains. This could be a research direction for the future.

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