Hilbert Transform Applications in Asynchronous Demodulation for Real Zero Single Sideband Signals in Mobile Radio Path

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Abstract This paper presents asynchronous demodulation methods without the threshold effect for a single sideband (SSB) with reduced carrier signal propagating through mobile radio paths. After an SSB with reduced carrier signal is converted into real zero SSB (RZ SSB) signals in a receiver, the relation of a Hilbert transform pair plays the main role in obtaining its practically perfect demodulation performance using DSP (Digital Signal Processing) processors. The asynchronous demodulation method, first developed in this study, is indispensable for supporting ‘burst mode transmission’ using SSB technology for voice/digital data communications with an ITU-T (International Telecommunication Union Telecommunication Standardization Sector) voiceband modem (MODulator-DEModulator).

Keywords: single sideband with reduced carrier, RZ SSB, asynchronous demodulation, flat Rayleigh fading, mobile radio path, Hilbert transform, DSP processor

1. Introduction

Single sideband (SSB) modulation [1] is always an attractive modulation in view of spectrum efficiency. However, it is difficult to utilize SSB modulation for land as well as maritime mobile radios in severe fading environments without implementing additional techniques to combat fading. The conventional method of correcting real or imaginary envelope fading is to transmit a pilot or carrier that is used for gain and/or phase control at the receiver. However, it was theoretically and experimentally revealed in a previous study [2] that an automatic gain control (AGC) circuit cannot cope with severe fading at all. Therefore, we have decided to pursue an asynchronous demodulation method, where we use neither AGC nor automatic frequency control (AFC) including a phase lock loop (PLL) circuit in the receiver to compensate for degradation by fading. This is very important to support ‘burst mode transmission’ for voice/data communications.

We have developed asynchronous demodulation methods for SSB signals in two phases. The first phase included the development of a real zero single sideband (RZ SSB) receiver, which we used to study the waveform of an SSB signal described by only real zeros [3]. Its theoretical formulation is based on an earlier paper by Logan [4]. To verify the performance of the RZ SSB receiver, we first fabricated RZ SSB demodulation circuits using discrete analog components for laboratory use and obtained promising results. Then, we fabricated RZ SSB transceivers to investigate in detail the performance of the proposed technique through indoor and outdoor tests, and confirmed that the RZ SSB receiver can support both voice and/or digital data communications with the severe fading encountered in VHF band land mobile radio paths [5], [6].

In the second phase, we implemented the RZ SSB receiver based on DSP processors. This is because the threshold effect observed in the SINAD (SInal plus Noise And Distortion to noise plus distortion ratio) performance [5], which arises with the introduction of an amplitude limiter just like in a frequency modulation (FM) receiver [1], cannot be removed from the performance of an RZ SSB receiver based on discrete analog components. The technique for overcoming this threshold effect and the fabrication of the entire demodulation circuit for the RZ SSB receiver are precisely demonstrated using the relation of a Hilbert transform pair [7] in section 3.

The rapid development of asynchronous demodulation methods for SSB signals using DSP processors is attributed to the need for various services using SSB modulation, which may become obsolete with the limitation of analog technologies. This study provides excellent tools for the implementation of SSB schemes in fading environments, especially for burst mode transmission, and in HF/VHF band mobile communications, tactical communications [8], [9],
underwater communications [10], [11], and so on.

2. Overview of RZ SSB Receiver Based on Discrete Analog Components

2.1 Objective of RZ SSB scheme

The RZ SSB scheme employs a reduced carrier USB as a transmitted signal because the carrier component is independent of the baseband signal and represents wasted power. However, the reduced carrier component is still useful not only for demodulating SSB signals but also for providing the information of various parameters such as received signal strength indication (RSSI), residual frequency offset, and so on. The level of the reduced carrier component is determined from the trade-off between the power efficiency and the performance of the demodulated signal. Consequently, we observed that the reduced carrier level of the RZ SSB signal is -13.2dB below the peak envelope power (PEP) [12], which is identical to the reduced carrier power of -11dB relative to the signal.

Figures 1 and 2 show a schematic diagram of an RZ SSB receiver without diversity reception and demodulation circuits based on discrete analog components, respectively. The RZ SSB receiver is composed of three parts: a linear amplification part, an RZ SSB signal generation part, and an RZ SSB signal demodulation part. In the RZ SSB generation part, the reduced carrier upper sideband (USB) is converted to the full-carrier lower sideband (LSB). During this conversion, the selective amplification of the reduced carrier component and the cancellation of random phase noise induced by fading are simultaneously performed in the intermediate frequency (IF) band [14].

2.2 Asynchronous demodulation

The input signal for the RZ SSB demodulation circuit shown in Fig. 2 can be described as

\[ r_d(t) = \alpha(t)(1 + mg_d(t)) \cos \omega_d t + mH[g_d(t)] \sin \omega_d t \]  

where \( \alpha(t) \) is the angular frequency of the IF band, \( \alpha(t) \) is the amplitude variation caused by flat Rayleigh fading, \( g_d(t) \) is a band-limited baseband signal, and \( m \) is the modulation index. When the arctangent nonlinearity of the arctangent function. When the modulation index is less than unity, the arctangent function \( \phi_d(t) \) can be expanded with respect to \( m \). In an earlier study [15], we derived the governing equation of the linearizer, which can cancel out the distortions up to the third order of \( m \). Applying a brute force method to

\[ r_d(t) = A_d(t) \cos(\omega_d t - \phi_d(t)) \]  

where

\[ A_d(t) = \alpha(t) \sqrt{(1 + mg_d(t))^2 + m^2 H^2[g_d(t)]} \]  

\[ \phi_d(t) = \arctan(mH[g_d(t)]/(1 + mg_d(t))) \]

Fig. 2 RZ SSB demodulation circuits with discrete analog components

\( H[g_d(t)] \) is the Hilbert transform of \( g_d(t) \). As the absolute maximum value of \( g_d(t) \) is normalized to unity in Eq. (1), \( m \) is the modulation index, which is less than unity. The random phase noise is already removed in the second part of Fig. 1. The rectangular form of Eq. (1) can be converted to the polar form as

\[ r_d(t) = A_d(t) \cos(\omega_d t - \phi_d(t)) \]  

where

\[ A_d(t) = \alpha(t) \sqrt{(1 + mg_d(t))^2 + m^2 H^2[g_d(t)]} \]  

\[ \phi_d(t) = \arctan(mH[g_d(t)]/(1 + mg_d(t))) \]

The baseband signal of the RZ SSB system based on discrete analog components is detected from the arctangent in Eq. (3), which implies that the carrier component in the demodulator is always unity. Consequently, asynchronous detection can be achieved.

The overall amplitude variation of the input signal in (2) is eliminated using the amplitude limiter in Fig. 2 to avoid unnecessary amplitude modulation to phase modulation (AM-PM) conversion. A ceramic frequency discriminator is used to obtain the derivative of the input signal, i.e., \( d\phi_d(t)/dt \). A -6dB/oct. equalizer is used to integrate the input signal, i.e., to obtain \( \phi_d(t) \). This is identical to Eq. (3), which enables the linearizer to eliminate distortions.

2.3 General governing equation of the linearizer

The linearizer is one of the key components of an RZ SSB receiver based on discrete analog components, which is used to cancel out the distortions generated by the nonlinearity of the arctangent function. When the modulation index is less than unity, the arctangent function \( \phi_d(t) \) can be expanded with respect to \( m \). In an earlier study [15], we derived the governing equation of the linearizer, which can cancel out the distortions up to the third order of \( m \). Applying a brute force method to
carefully identify the distortion generation mechanism, we can heuristically construct the general governing equation of the linearizer \( u(t) \) as follows:

\[
u(t) = \sum_{j=0}^{\infty} \left( \frac{1}{(2j+1)!!} \right) ^ 2 C_2^{2j} \phi_{d,2j}(t) H_{2j}^{2j+1}(\phi_{d}(t))
\]

Here, \( H[\phi_{d}(t)] \) is the Hilbert transform of \( \phi_{d}(t) \) and \( C_2^{2j} \) is the coefficient of the binomial expansion. The cancellation capability of the linearizer is investigated using a DSP based spectrum analyzer in the next section.

Taking the limit of Eq. (4) at infinity, we get [15]

\[
limit_{j \to \infty} u(t) = m H[g_n(t)]
\]

In fact, the simulation results in Fig. 3 show that the above relation holds for even \( j=4 \) (corresponding to linearization up to 7th and 8th order distortions) in practical situations.

Next, we discuss the difference between \( g_n(t) \) and \( H[g_n(t)] \). In the case of voice/speech/music sounds, human ears cannot distinguish between the two terms because the power spectrum of \( g_n(t) \) is identical to that of \( H[g_n(t)] \). When the voiceband modem (Modulator-DEModulator) signal is considered, its modulated signal such as the quadrature amplitude modulation (QAM) scheme is generally described as

\[
m_q(t) = s_i(t) \cos \omega_0 t - s_q(t) \sin \omega_0 t
\]

where \( s_i(t) \) and \( s_q(t) \) are the in-phase and quadrature baseband digital signals, respectively. These are used to form a constellation diagram. \( \omega_0 = 2 \pi f_0 \) is the angular frequency, for example, \( f_0 = 1700 \) Hz for ITU-T (International Telecommunication Union Telecommunication Standardization Sector) V.29 and \( f_0 = 1800 \) Hz for ITU-T V.32 bis. The Hilbert transform of Eq. (6) can be obtained using Bedrosian’s theorem [7] as

\[
H[m_q(t)] = s_i(t) \sin \omega_0 t + s_q(t) \cos \omega_0 t
\]

Subsequently, the mathematical relation between (6) and Eq. (7) can be obtained as follows:

\[
m_q(t) = m_q(t) + j H[m_q(t)]
\]

\[
= (s_i(t) + js_q(t)) \exp(j \omega_0 t)
\]

Therefore, the bit error rate (BER) remains the same irrespective of whether it is measured using Eqs. (6) or (7). However, if the original waveform at the receiving point is required, \( m H[g_n(t)] \) of Eq. (5) should be converted to \( m g_n(t) \).

### 2.4 Performance of linearizer

As the general governing equation of the linearizer, i.e., Eq. (4), is heuristically derived, it is necessary to demonstrate the cancellation performance of the linearizer using MATLAB. To this end, we use a square wave such as a two-tone signal as the test signal [16]

\[
f(t) = \cos(2 \pi f t) - (1/3) \cos(6 \pi f t)
\]

The absolute maximum value of \( f(t) \) is 0.943. When \( f(t) = m g_n(t) \), the modulation index becomes \( m = 0.943 \).

Figure 3(a) illustrates the spectrum of the input signal for the linearizer, which is given by Eq. (8), where the frequency of the two-tone signal is set to be 300 and 900Hz for convenience. As the modulation index approaches unity, the number of harmonic components increases. In this simulation, we conveniently consider a high sampling frequency of 32kHz to clearly observe the harmonic components without folded error distortions. Figures 3(b)–3(f) indicate the spectra in which the distortions up to the 5th, 6th, 7th, 8th, and 9th order are canceled out by the linearizer, respectively. Although the modulation index of 0.943 generates the higher order distortions, the linearizer facilitates the rapid cancellation of these distortions. The results obtained by the DSP based spectrum analyzer validate the effectiveness of the linearizer in obtaining the demodulated baseband signal.

### 2.5 Achievements of RZ SSB transceivers based on discrete analog components

This section provides a brief review of the achievements of RZ SSB transceivers in land and maritime mobile radio environments within the limitations of analog devices.

a. The RZ SSB transceiver is the first SSB technology in the world that utilizes neither AGC nor AFC/PLL circuits to combat severe fading. In voice/speech communications, we do not hear the ‘Donald Duck’ voice generated by a mistuned carrier in the audio signal output [3], [5], [6], [15].

b. Taking advantage of the burst mode transmission of an RZ SSB transceiver, we can utilize the fall back mode and automatic repeat request (ARQ) to improve communication quality. JPEG (Joint Photographic Expert Group)/text files have been transmitted between two coastal stations approximately 2000 km apart through an HF radio path by switching between transmission and reception just like in ping-pong transmission [17], even before the ITU-T recommendation for the JPEG color facsimile was
c. It is essential to manage the maximum value of a baseband signal. In particular, human voices possess a wide dynamic range; therefore, we have implemented a control circuit with a practical dynamic range [19].

d. Two branch space diversity reception with an equal gain combiner is useful for improving the quality of received signals such as voice/digital data [5], [14].

e. The introduction of a circuit for eliminating random phase noise is effective even for a single branch receiver [14].

f. We obtained excellent performance for both voice and digital data transmission above the threshold. However, unfortunately we cannot overcome the threshold effect in an RZ SSB receiver based on discrete analog components [5], [15], which arises with the introduction of an amplitude limiter. This issue is resolved in an RZ SSB transceiver based on DSP processors.

3. DSP Processor Based RZ SSB

3.1 DSP based approach

First, we provide a brief description of the analytic signal in polar coordinates, which is expressed as

\[ z(t) = A_s(t) \exp(j\phi_s(t)) \]  

As the envelope of \( A_s(t) \) is always positive for a full-carrier SSB, let us consider the natural logarithm of Eq. (9).

\[ \ln z(t) = \ln A_s(t) + j\phi_s(t). \]  

The real and imaginary parts of (10) are mutually related by a Hilbert transform pair [7]. Consequently, we have

\[ \phi_s(t) = H[\ln A_s(t)] \]  
\[ \ln A_s(t) = -H[\phi_s(t)] \]  

From Eq. (9), the polar coordinates \((A_s(t), \phi_s(t))\) are converted to the rectangular coordinates \((x_s(t), y_s(t))\) as

\[ x_s(t) = A_s(t) \cos \phi_s(t) \]  
\[ y_s(t) = A_s(t) \sin \phi_s(t) \]  

In RZ SSB based on discrete analog components, the linearizer is used for obtaining a distortion less baseband signal from the instantaneous phase angle in Eq. (3). In the DSP processor based RZ SSB, instead of the linearizer, we can accurately obtain the amplitude of \( A_s(t) \) through the Hilbert transform of \( \phi_s(t) \), i.e., from Eq. (12), after calculating the phase angle \( \phi_s(t) \) from the received signal. Subsequently, the baseband signals are obtained from Eqs. (13) and (14).

Fig. 3 Effectiveness of linearizer in canceling out distortions, where (a) is the spectrum of the input signal and (b)-(f) are output spectra of the linearizer.
3.2 Transmitted signal

The transmitted signal of a reduced carrier USB can be described in terms of a complex envelope [20] as follows:

\[ s(t) = k_c + g(t) + jH[g(t)] \quad (15) \]

Here, \( g(t) \) is a band-limited baseband signal and \( H[g(t)] \) is the Hilbert transform of \( g(t) \). The maximum absolute value of \( g(t) \) is assumed to be unity. \( k_c \) is a DC component corresponding to the level of the reduced carrier component.

3.3 Fading channel

Figure 4 illustrates a schematic diagram of the propagation path from the RZ SSB transmitter to the receiver. For convenience, the complex baseband signals are illustrated with double lines. When the transmitted signal is fed into a fading environment, it is severely affected by multiplicative fading noise. Furthermore, thermal noise is added to the signal at the receiver. Therefore, the received signal is given by

\[ r(t) = c(t)s(t) + n(t) \quad (16) \]

The term \( c(t) \), which also appeared in Eq. (1), is statistically governed by the Rayleigh distribution. The function \( c(t) \) is described by the uniform distribution, i.e., random phase noise, which did not appear in Eq. (1). This is because Eq. (1) corresponds to the case in which the random phase noise is already eliminated by the noise removing circuit [14] in the second part of Fig. 1.

where \( n(t) \) is additive white Gaussian noise (AWGN) with the power spectral density \( n_0 \) in both the real and imaginary components. The complex gain \( c(t) \) of the channel incorporates both the frequency offset and fading noises [21] as

\[ c(t) = g_c(t) \exp j\Delta f t \quad (17) \]

where \( \Delta f_0 = 2\pi f_0 \) and \( f_0 \) is the residual frequency offset, which is determined from the frequency stability of oscillators installed in both the transmitter and the receiver. The power spectrum of the complex gain is described by a U-shaped spectrum, whose central frequency in this model is \( f_0 \). The overall bandwidth is twice the maximum Doppler frequency \( f_d \), which is calculated as \( f_d = 9.25 \times 10^3/v \), where \( f \) is the radio frequency (RF) in MHz and \( v \) is the maximum vehicular speed in km/h [2].

On the basis of the statistical model of flat Rayleigh fading and the implementation of its simulator [2], the time function \( g_c(t) \) can be expressed as

\[ g_c(t) = \alpha(t) \exp j\phi(t) \quad (18) \]

The term \( \alpha(t) \), which also appeared in Eq. (1), is statistically governed by the Rayleigh distribution. The function \( \phi(t) \) is described by the uniform distribution, i.e., random phase noise, which did not appear in Eq. (1). This is because Eq. (1) corresponds to the case in which the random phase noise is already eliminated by the noise removing circuit [14] in the second part of Fig. 1.

3.4 Receiver signal processing

Figure 5 shows a detailed block diagram of the receiver. For simplicity, the following signal processing is only described for single branch reception. We first
present a practical approach to searching for an asynchronous demodulation method for RZ SSB signals under the assumption that the power of $c(t) s(t)$ is much larger than that of $n(t)$. For convenience when fabricating RZ SSB transceivers equipped with the algorithms newly developed in this section, we will experimentally examine in the following section whether

Using Eqs. (15)-(18), we get the received signal $r(t)$

$$r(t) = a(t)(k_c + g(t) + jH[g(t)])\exp j(\Delta_0t + \varphi(t))$$  \hspace{1cm} (19)

Here, we consider that a finite impulse response low-pass filter (FIR-LPF), which exhibits a linear phase response, is important for the signal processing. We should pay attention to the passband ripple on the narrowband FIR-LPF because the phase component as possible must be extracted from Eq. (23). Equation (21) is not affected by the passband ripple of the narrowband FIR-LPF, whose passband peak-to-peak ripple $(2\delta)$ is approximately $2\delta=0.2\text{dB}$. The frequency band edges of the FIR-LPF should be determined to evaluate the frequency stability of oscillators installed in the transmitter and receiver and the spread of the Doppler spectrum.

The reduced carrier component located near zero frequency in the receiver is extracted using the narrowband FIR-LPF from Eq. (19), i.e.,

$$r_{\alpha}(t) = k_c a(t) \exp j(\Delta_0t + \varphi(t))$$  \hspace{1cm} (20)

Here, $a(t)$ denotes the output of the narrowband FIR-LPF affected by the passband ripple. The phase angle term of Eq. (20) can be calculated as

$$r_{\alpha}(t) = r_{\alpha'}(t)/\sqrt{r_{\alpha'}(t)r_{\alpha'}^*(t)}$$

$$= \exp j(\Delta_0t + \varphi(t))$$  \hspace{1cm} (21)

where $y^*(t)$ denotes the complex conjugate of $y(t)$. Equation (21) is not affected by the passband ripple of the narrowband frequency-demodulation component because the phase component is perpendicular to the amplitude component. The amplitude term of Eq. (20) can be calculated as

$$r_{\alpha}(t) = \sqrt{r_{\alpha'}(t)r_{\alpha'}^*(t)}$$

$$= k_c a(t)$$  \hspace{1cm} (22)

which is the degenerated carrier component.

Delay line #1 in Fig. 5 is introduced for the input signal of $r(t)$ to compensate for the delay time introduced by the narrowband FIR-LPF. For simplicity, we neglect the delay time associated with the narrowband FIR-LPF in the mathematical formulation. To remove the random phase noise from Eq. (19), we multiply Eq. (19) by the complex conjugate of Eq. (21), i.e.,

$$r_{\alpha}(t) = r(t)r_{\alpha'}^*(t)$$

$$= a(t)(k_c + g(t) + jH[g(t)])$$  \hspace{1cm} (23)

During this process, we can perform both the cancelation of the random phase noise caused by fading and a homodyne detection. This process for the receiver based on the discrete analog components is carried out in the IF band.

As the reduced carrier component is unnecessary in the following signal processing, as much of the component as possible must be extracted from Eq. (23) using Eq. (22). To complete the signal processing, the passband ripple on the narrowband FIR-LPF should be negligible. The fabricated narrowband FIR-LPF exhibits ripple of $2\delta=0.2\text{dB}$; however, we employ it for the demodulation process under the assumption that $\alpha(t)=a(t)$. It is necessary to verify the feasibility of this assumption by investigating the SINAD and BER performance characteristics through indoor experiments on the RZ SSB receiver. Subsequently, the reduced carrier component in Eq. (23) is substituted using Eq. (22) under the above assumption to obtain

$$r_{\alpha}(t) = a(t)(g(t) + jH[g(t)])$$  \hspace{1cm} (24)

We now describe the demodulation process in the DSP processor using Eqs. (12), (22), and (24). Firstly, considering the performance of the FIR Hilbert transformer, we shift the frequency position of (24) using the oscillator output $\exp(-j\omega_0t)$, whose frequency does not overlap with the spectra of $g(t)$, i.e.,

$$r_{\alpha}(t) = a(t)(g(t) + jH[g(t)])\exp j(-\omega_0t)$$  \hspace{1cm} (25)

The degenerated carrier component in Eq. (22) is amplified up to $K_a(t)$, where the multiplication factor is $M_a=K/k_c$. To apply Eq. (12), the amplified DC component is added to Eq. (25) to generate the full-carrier SSB signal. $K$ must satisfy the full-carrier constraint even though both $g(t)$ and $H[g(t)]$ are frequency-shifted. Consequently, we have

$$r_{\alpha}(t) = K_a(t) + a(t)(g(t) + jH[g(t)])\exp j(-\omega_0t)$$

$$= a(t)(K + g(t)\cos \omega_0t + H[g(t)]\sin \omega_0t)$$

$$+ ja(t)(-g(t)\sin \omega_0t + H[g(t)]\cos \omega_0t)$$  \hspace{1cm} (26)

The instantaneous phase angle of $\phi(t)$ in Eq. (26) can be calculated using the arctangent as
\[
\phi(t) = \arctan \frac{\alpha_r(t)(-g(t)\sin \omega_0 t + H[g(t)]\cos \omega_0 t)}{\alpha_r(t)(K + g(t)\cos \omega_0 t + H[g(t)]\sin \omega_0 t)}
\] (27)

As the term of \( \alpha_r(t) \) in Eq. (27) is common to both the denominator and the numerator, it can be completely canceled out. Therefore, the functioning of an amplitude limiter can be mathematically introduced into the signal processing.

To obtain the amplitude component of \( A(t) \) using Eq. (12), the function of \( \phi(t) \), which is non-bandlimited, must be Hilbert-transformed. An FIR Hilbert transformer exhibiting a wide passband and a small ripple is required for this process. Consequently, it is convenient to shift the full-carrier SSB signal of Eq. (26) to an appropriate frequency on the Hilbert transformer. In the DSP processor, we have implemented FIR Hilbert transformer \#1, whose peak-to-peak ripple is less than 0.01dB, and the passband is wide enough to avoid unnecessary folded errors generated by the non-bandlimited function of \( \phi(t) \) and to ensure excellent SINAD characteristics. This is the reason for the requirement of the frequency shift.

Referring to Fig. 5 and Eq. (12), the desired amplitude component of \( A(t) \) is obtained as

\[
A(t) = \exp(-H[\phi(t)])
\] (28)

After calculating \( \sin \phi(t) \) using Eq. (27), we get

\[
r_{\text{os}}(t) = A(t)\sin \phi(t) \\
= -g(t)\sin \omega_0 t + H[g(t)]\cos \omega_0 t \tag{29}
\]

where the second line is obtained from Eq. (14). During these processes, as FIR Hilbert transformer \#1 is used, its delay time must be compensated by delay line \#2. For simplicity, we do not introduce a time delay into the mathematical formulation.

To shift the frequency of the baseband signal in Eq. (29) back to its original frequency position, an analytic signal formulation is convenient for DSP processors.

Using delay \#3 and Hilbert transformer \#2 in Fig. 5, we get the following equation:

\[
r_{\text{os}}(t) = r_{\text{os}}(t) + jH[r_{\text{os}}(t)] \\
= -g(t)\sin \omega_0 t + H[g(t)]\cos \omega_0 t \\
+ jg(t)\cos \omega_0 t + H[g(t)]\sin \omega_0 t \\
= H[g(t)]\exp(j\omega_0 t)
\] (30)

The delay time for Hilbert transformer \#2 is also neglected in the above formulation. Multiplying Eq. (30) by the oscillator output signal of \( -e^{j\omega_0 t} \) adjusted by the phase shifter in Fig. 5, we can get the demodulated baseband signal in complex form as

\[
r_{\text{dem}}(t) = g(t) - H[g(t)]
\]

We can choose the real part of the above equation, which is identical to the transmitted baseband signal. Here, the multiplicative noise due to fading is completely eliminated.

4. Experimental Results

4.1 Indoor experimental setup

Figure 6 shows the indoor experimental setup for measuring SINAD or BER characteristics. Experimental results are useful for demonstrating the performance of the RZ SSB transceiver, investigating whether the threshold effect is removed, and examining the feasibility of the assumption for passband ripple. The RZ SSB transceivers governed by ARIB (Association of Radio Industries and Businesses) STD (Standard)-T62 [12] and ARIB TR (Technical Report)-B21 [13], whose radio frequency is the 165MHz band, are fabricated using DSP processors with the new software described in this paper. As the RZ SSB receivers are equipped with two branch space diversity reception with an equal gain combiner (2-BSD/EGC) [2], we have installed two flat Rayleigh fading simulators (product of JRC) into the setup.
4.2 SINAD performance

Figure 7 depicts the relation between SINAD and receiver average input level, where open circles represent the data measured under AWGN. Triangles and crosses indicate the data for 20Hz flat Rayleigh fading with diversity and without diversity, respectively. The dotted line along the circles intersects the abscissa at approximately -21.5dBμV, which corresponds to the receiver sensitivity $P_s$ defined as $P_s = kTBF/G$ [1]. The receiver sensitivity for the present receiver of -21.5dBμV can be easily calculated by introducing the common logarithm, $kT$ of -173.8dBm/Hz ($k$: Boltzmann constant and $T$: absolute temperature of 300K), plus the noise bandwidth $B$ of 35.3dB (=10log3400), plus the noise figure $F$ of 7dB, minus the process gain of the equal gain combiner, $G$, of 3dB. Note that 1μV (=0dBμV), which corresponds to an open circuit, is equivalent to the power level of -113dBm for 50Ω source resistance.

When the radio frequency is 165MHz, 20Hz fading is obtained at a vehicular speed of 131km/h. The experimental SINAD characteristics in Fig. 7 show that no threshold exists at a low input, i.e., less than approximately -10dBμV in AWGN environments, whereas the threshold effect is always observed in this region in the RZ SSB receivers based on discrete analog components [5], [15]. SINAD is proportional to the input level of the receiver. However, a saturation behavior is observed at a relatively high input level, which is identical to that observed in the RZ SSB receiver based on discrete analog components. Therefore, the saturation level has the same origin in the two RZ SSB receivers, that is, some distortions generated from the linear amplification part may be one of the causes of the saturation. We observed that the saturation causes neither the ripple effect of the narrowband FIR-LPF nor the quantization noise due to the finite word length of the DSP. SINAD characteristics in fading environments are effectively improved in the absence of a threshold effect. Furthermore, we did not observe any degradation in SINAD due to the passband ripple of the narrowband FIR-LPF.

4.3 BER performance

We performed transmission experiments using voiceband modems [22] such as ITU-T V.29 16QAM with a data transmission rate of 9.6kbps, ITU-T V.32 bis 64/128QAM with that of 14.4kbps, and V.32 bis 256QAM with that of 19.2kbps (modems are products of NEC). The RZ SSB transceiver is only installed in the forward link and a metallic wire is installed in the backward link.

Figure 8 shows the relation between bit error rate (BER) and bit energy-to-noise density ratio (Eb/No) for 16QAM/9.6kbps on the RZ SSB with the reduced carrier power of -11dB. Here, open and closed circles denote data for AWGN and 20Hz flat Rayleigh fading, respectively. The RZ SSB receiver is equipped with 2-BSD/EGC. For later discussion, the BER performance for ITU-T V.29 16QAM/9.6kbps on a transparent tone-in-band (TTIB)-SSB scheme [23] is depicted in the same figure. Here, open and closed squares denote data for AWGN with pilot powers of -3dB and -9dB, respectively. Note that the BER data for TTIB-SSB is reproduced from Fig. 7 depicted in [24].

The measured BER data on the RZ SSB for 16QAM under AWGN is in good agreement with the theoretical curve in Fig. A-1 (see Appendix), although a 2.5dB degradation is observed because commercial modem products are equipped with a waveform-shaping filter. A dotted line is drawn along the BER data for 20Hz flat Rayleigh fading in Fig. 8, whose $\Gamma^2$ dependence is
The dotted line is added to express that SINAD is proportional to the input level of the receiver. However, a saturation effect is observed. Figure 7 depicts the relation between SINAD and input level of 35.3dB (=10log3400), plus the logarithm of the noise bandwidth. The measured BER data on RZ SSB for 16QAM, 64QAM(256QAM, respectively. where BER is depicted under AWGN and 20Hz fading environments. Note that the receiver is equipped with 2-BSD/EGC.

Theoretical data on BER for AWGN in Fig. 9 deviates from the theoretical characteristics in Fig. A-1, especially for 64QAM and 256QAM. Estimating graphically the $\Gamma^n$ dependence from the BER data for fading in Fig. 9, we obtain $n=1.7$, 1.6, and 1.34 for 128QAM (TCM), 64QAM, and 256QAM, respectively. There is a discrepancy between these estimated values and the theoretical value of $n=2$. These two issues require further study.

Although ITU-T voiceband modems with 16/64/128/256QAM schemes were developed for AWGN switched telephone networks, the present experimental results indicate that RZ SSB transceivers can support the transmission of ITU-T 16/64/128/256QAM schemes even in land mobile radio environments.

5. Simulation Results and Discussion

5.1 Role of degenerated carrier in demodulation process

To confirm that the threshold effect does not exist in Fig. 7, we executed computer simulations for the following two cases using MATLAB: a degenerated carrier and a conventional carrier. For convenience, we assume that residual frequency offset and fading do not exist, but thermal noise does. The received SSB with the reduced carrier signal at the input of the receiver can be described as

$$r_n(t) = k_c + g(t) + j\|g(t)\| + n_r(t) + jn_q(t)$$  \hspace{1cm} (31)

The in-phase and quadrature noise terms, $n_r(t)$ and $n_q(t)$, respectively, are thermal noise variables with the power spectral density $n_v$. The reduced carrier $k_c$, which is contaminated by thermal noise, is first extracted from the narrowband FIR-LPF. This signal is expressed as

$$r_n(t) = k_c + n_{r_n}(t) + jn_{q_n}(t)$$  \hspace{1cm} (32)

Here, $n_{r_n}(t)$ and $n_{q_n}(t)$ are the thermal noises associated with the reduced carrier signal, whose noise bandwidth $B_n$ is twice the bandwidth of the narrowband FIR-LPF.
The reduced carrier \( k_c \) in Eq. (31) is replaced by the degenerated carrier in Eq. (33), whose level is large enough to satisfy the condition of a full-carrier SSB. The SSB signal with the degenerated carrier-plus-noise is given as

\[
r_{dg}(t) = K_{dg} + (g(t) + n_c(t)) + j[H[g(t)] + n_r(t)]
\]

Both \( g(t) \) and \( H[g(t)] \) are only affected by \( n_c(t) \) and \( n_r(t) \), respectively. The noise bandwidth of \( B \) is equal to that of the SSB signal. The instantaneous phase angle in Eq. (34) is expressed as

\[
\phi_{dg}(t) = \arctan \frac{(H[g(t)] + n_r(t))/(K_{dg} + g(t) + n_c(t))}\]

Let the denominator and numerator in the fraction of the arctangent function in Eq. (35) be \( x(t) \) and \( y(t) \), respectively. We have

\[
x(t) = K_{dg} + (g(t) + n_c(t))
\]
\[
y(t) = H[g(t)] + n_r(t)
\]

The locus generated by the point \( (x(t), y(t)) \) can be depicted on the \( x-y \) plane through the time functions \( g(t) \) and \( H[g(t)] \). We can simulate the locus of the SSB signal with the degenerated carrier using the test signal in (8), where the frequency of the two-tone signal is set to be 300 and 900 Hz for convenience.

The simulation is performed, for example, for \( \text{SINAD}=5 \) dB, where the threshold effect always appeared in earlier studies [5], [15]. Figure 10 (a) depicts the simulation result, where the locus is confined to the 1st and/or 4th quadrant. This result practically guarantees that there is no threshold effect as shown in Fig. 7.

It is useful to simulate a similar locus under \( \text{SNR}=5 \) dB for the conventional full-carrier SSB signal. In this case, the level of the extracted carrier is selectively increased by a factor of \( M_A \). The simulation result for the conventional full-carrier SSB is depicted in Fig. 10 (b), where the locus is not confined to the 1st and/or 4th quadrant. When the locus crosses over the \( y \)-axis, i.e., \( x=0 \), a sudden \( 2\pi \) radians phase excursion occurs. This sudden phase change makes a “click” [1], which significantly decreases the SINAD of the demodulated signal. Both positive and negative \( 2\pi \) radians phase excursions can occur with equal probability. The probability of such a phase excursion increases as \( \text{SNR} \) decreases to a value below 9 dB. It is clarified that the degenerated carrier is very important for overcoming the threshold effect during the RZ SSB demodulation.

5.2 Synchronous demodulation for SSB

5.2.1 TTIB-SSB scheme

It was theoretically pointed out in [25] that the original filter employed in the phase-locked TTIB (PLTTIB)-SSB scheme [23], [26] is not appropriate for generating the spectral gap in which the pilot signal is placed. Thus, the original filter was replaced with an appropriate filter, the quadrature mirror filter (QMF), to eliminate the self-noise [27]. The analysis of PLTTIB-SSB with a symmetric phase detector in [28] showed that the phase jitter can be reduced by narrowing the bandwidth of the tracking filter, although this increases the acquisition time.

During the field test on the GMSK scheme [29], we
encountered the same difficulties in finding an optimal trade-off between phase jitter and acquisition time [30]. Thus, we were motivated to search for asynchronous demodulation methods for an RZ SSB signal.

The BER performance with TTIB-SSB assistance was measured for ITU-T V.29 16QAM/16kbps with a roll-off factor of 0.3 in [24]. As the pilot power level decreased from -3 to -9dB, the discrepancy from the theoretical curve increased from 1 to 5dB when the pilot energy was excluded for Eb/No [24].

Figure 8 shows the BER performance characteristics for two schemes, where open and closed rectangles denote BER data on TTIB-SSB for AWGN with pilot powers of -3 and -9dB, respectively. Note that the BER data was obtained with the pilot energy for Eb/No included. The BER data on TTIB-SSB with the pilot power of -9dB [24] is around 2dB near BER=10⁻³, inferior to that of RZ SSB. However, TTIB-SSB with the pilot power of -9dB [24] is around 2dB near BER=10⁻³, inferior to that of RZ SSB, even though RZ SSB takes the reduced carrier power of -11dB. It is generally considered that the BER performance under AWGN employed for synchronous demodulation is superior to that for asynchronous demodulation. However, this consideration is not applicable to the TTIB-SSB (see Fig. 8). Note that there are no technical reports on 64/128/256QAM performance and/or on SINAD characteristics with the revised TTIB-SSB scheme.

5.2.2 Frequency demodulation

A frequency demodulation method was proposed in [31], where a received SSB with a certain level of carrier signal was effectively resolved into two SSB signals, with the carrier levels of each appropriately manipulated. Then, they were frequency-demodulated to completely cancel out the even order distortions generated by the frequency demodulation process. Moreover, the third order distortion can be easily reduced without using Hilbert transformer. It was claimed that the method can be applied to the RZ SSB without using a linearizer. However, the performance under fading environments as well as AWGN using practical voice/digital data signals was not demonstrated. Anyway, to execute the proposed demodulation process, the regeneration of the carrier from the received SSB signal is required.

5.3 Asynchronous modulation for SSB

Before the advent of DSP technology, Voelcker and Peoples published technical papers on the envelope detection of full-carrier SSB signals [32], [33]. Their method can be only applied to full-carrier SSB under non-fading.

Peoples obtained a U.S. patent [34] for his invented asynchronous demodulation method using an instantaneous phase term just like in this study. As his model took the conventional full-carrier SSB, it is easy to conceive that his method generates the threshold effect explained by Fig. 10(b). He also did not take any effective measure for severe fading.

5.4 Alternative demodulation methods

The following alternative methods have been mathematically derived, although it is necessary to evaluate their practical performance characteristics in detail.

5.4.1 Alternative method #1

Using Eqs. (13), (27), and (28), we can derive the following equation:

\[ r_c(t) = A(t) \cos \phi(t) \]
\[ = K + g(t) \cos \omega_c t + H[g(t)] \sin \omega_c t \] (36)

Removing the DC component, i.e., \( K \), from Eq. (36), we get

\[ r_{\text{cal}}(t) = g(t) \cos \omega_c t + H[g(t)] \sin \omega_c t \] (37)

The Hilbert transform [7] of Eq. (37) is

\[ H[r_{\text{cal}}(t)] = g(t) \sin \omega_c t - H[g(t)] \cos \omega_c t \] (38)

Equations (37) and (38) can be used to generate the analytic signal as

\[ r_{\text{cal}}(t) = r_{\text{cal}}(t) + jH[r_{\text{cal}}(t)] \]
\[ = (g(t) - jH[g(t)]) \exp(j \omega_c t) \] (39)

When the term \( \exp(j \omega_c t) \) in Eq. (39) is removed from the output of the oscillator, we get the final demodulated baseband signal as

\[ r_{\text{dom}}(t) = g(t) - jH[g(t)] \] (40)

The real part of Eq. (40) is identical to the transmitted baseband signal.

5.4.2 Alternative method #2

The analytic signal, whose real and imaginary parts are Eqs. (37) and (29), respectively, can be written as

\[ r_{\text{cal}}(t) = g(t) \cos \omega_c t + H[g(t)] \sin \omega_c t \]
\[ + j(-g(t) \sin \omega_c t + H[g(t)] \cos \omega_c t) \]
\[ = (g(t) + jH[g(t)]) \exp(-j \omega_c t) \] (41)

When the term \( \exp(-j \omega_c t) \) in Eq. (41) is removed from the conjugated oscillator output, i.e., \( \exp(j \omega_c t) \), we get the final demodulated baseband signal, i.e., \( g(t) + jH[g(t)] \).

In this method, delay #3 and Hilbert transformer #2 in
Fig. 5 are not required to obtain the final demodulated signal.

5.4.3 Alternative method #3

From Fig. 5, we can get the Hilbert transform pair, i.e., the output of delay #2 and Hilbert transformer #1. These two signals are then substituted into Eq. (4). Subsequently, we obtain the output of the linearizer, i.e., Eq. (29). After this calculation on the linearizer, the frequency shift process is required.

5.4.4 Alternative method #4

This is a simple and easy method. Dividing Eq. (24) by Eq. (22), we get

\[ r_w(t) = r_c(t)/r_w(t) = (g(t) + jH[g(t)]) / k_c \]

This method cannot be conceived without using the complex envelope of Eqs. (15) and (19).

5.5 Graceful degradation

Apart from that at the edge of the coverage area, the audio quality of RZ SSB for a voice signal is gracefully degraded due to the non-threshold, whereas that of a digital voice degrades quickly to digital chaos, where it is impossible to hear a normal voice. This is because the digital voice inherently requires a certain signal strength to secure a minimal BER and to avoid digital chaos. This phenomenon may be a type of threshold effect. Note that the level of the threshold in fading is much higher than that in thermal noise.

6. Conclusions

In this study, we demonstrated how to implement asynchronous demodulation using DSP processors into SSB receivers, which can support burst mode transmission. The SSB scheme has become an attractive modulation technology for narrowband mobile radios owing to the absence of cumbersome synchronous problems.

The threshold may occur at the SINAD point where the demodulated SINAD does not maintain a linear relation with the input signal level. Below the threshold point, the noise signal may instantaneously have an amplitude greater than that of the desired signal, thus the demodulated SINAD may be greatly degraded. The threshold can be successfully removed using DSP technologies but not the conventional technologies using discrete analog components. The computer simulations have clearly revealed the role of the carrier as indicated in Figs. 10(a) and 10(b), that is, the degenerated carrier component contaminated with thermal noise is important for removing the threshold effect in the RZ SSB receiver.

The general governing equation of the linearizer, which is one of the key devices, has been heuristically derived in this paper, and the effectiveness of the linearizer is demonstrated using MATLAB. The linearizer can cancel out the distortion faster than expected, although the modulation index has a value close to unity.

The proposed method, which plays an important role in the receiver configuration of an SSB operating in severe fading environments, is robust and versatile, and it facilitates the utilization of SSB modulation in VHF and HF mobile radios for land and maritime mobile communications.

Appendix

Theoretical examinations of BER performance for M-ary QAM schemes under AWGN and 2-BSD/EGC

The average probability of the error rate when employing Gray coding for square 16/64/256QAM under AWGN was given by using a complementary error function in [35] as

\[ P_{M-QAM}(\gamma) = A_1 \text{erfc}(\sqrt{\gamma}) - A_2 \text{erfc}^2(\sqrt{\gamma}) \]  (A-1)

where the numerical values for \(A_1, A_2,\) and \(B\) are tabulated in Table A-1, and \(\gamma\) is the instantaneous Eb/No. The theoretical relationship between BER and Eb/No (dB) for 16/64/256QAM are depicted in Fig. A-1.

To improve the quality of received signals, \(M\)-branch space diversity reception with equal gain combiner (\(M\)-BSD/EGC) techniques are introduced for flat Rayleigh fading environments. If the average Eb/No of \(\Gamma\) is the same for all the branches, the probability density function for \(M\)-BSD/EGC is given as [1]

\[ p(\gamma) = K_{EG} \left( \gamma^{M-1} / \Gamma^{M} \right) \]  (A-2)

where \(K_{EG} = M^{2M-1} M! / (2M-1)!\). BER characteristics are calculated using Eqs. (A-1) and (A-2) as

\[ P_{BER}(\Gamma) = \int_0^\infty P_{M-QAM}(\gamma) p(\gamma) d\gamma \]  (A-3)

Here, the second term of Eq. (A-1) may be neglected.
when $\gamma$ is much larger than unity, then, putting $(B_{\gamma})^{1/2} = x$, we can calculate the definite integral of Eq. (A-3) as follows:

$$P_{\text{BER}}(\Gamma) = \left(2A_{E_{G}}/\Gamma B_M\right)\int_0^\infty x^{-\frac{M-1}{2}} \text{erfc}(x) dx$$

$$= \left(A_{E_{G}}/\Gamma (M+1/2) / 2MB_M\right) / (1/\Gamma_M)$$

(A-4)

Here, $\Gamma(M+1/2)$ is a gamma function. When $M=2$, $\Gamma(5/2) = 3(\pi)^{1/2}/4$. The BER performance for 2-BSD/EGC is also depicted in Fig. A-1, where the pair of solid lines (left: BER for AWGN, right: BER for flat Rayleigh fading) represents 16QAM. Dashed and dash-dotted lines represent 64QAM and 256QAM, respectively.

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