Research Article

Improvement of College English Teaching Model under the Background of Information Technology

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In order to improve the effect of college English teaching reform, this paper combines information technology to improve the college English teaching mode. According to the digital processing requirements of English teaching resources, in order to make the loop establishment circuit time of automatic gain control independent of the input signal, this paper uses a logarithmic amplifier to detect it. Moreover, this paper combines the SC algorithm and the SC algorithm based on the time-domain PN sequence to propose a new frame synchronization algorithm. The algorithm has precise frame synchronization points and reduces the usage of hardware resources. Finally, this paper constructs a college English teaching system based on information technology. The experimental research results show that the college English teaching system based on information technology proposed in this paper has a good teaching effect.

1. Introduction

The evaluation index system of English teaching quality is an important determinant of whether the evaluation of college English teaching quality is scientific, fair, and reasonable. It is a very complicated issue to evaluate the quality of English teaching in higher education schools. Due to the differences in school types, course types, and student types, it is difficult to find a universally applicable English teaching quality evaluation system that can be used to evaluate teacher English teaching quality in all colleges and universities [1]. For different types of higher education schools, according to their own development and future goals, more targeted and more suitable evaluation indicators can be used to make the English teaching quality evaluation system more accurate and more scientific. This has a very important effect on the evaluation of English teaching quality [2]. The reasonableness of the comprehensive evaluation results of English teaching quality can enable school leaders, the staff of the English teaching quality evaluation management department, and the teachers themselves to understand and discover their own problems in English teaching in a timely manner and make appropriate adjustments in time. Moreover, it has an irreplaceable effect on the development of colleges and universities, the cultivation of students, and the continuous improvement of teachers’ English teaching level [3].

The English teaching quality evaluation index system should follow the following basic principles: according to the training objectives of college students, starting from the foundation of quality education, the construction of the English teaching quality evaluation system in this paper should follow the principle of comprehensiveness and examine it from multiple angles and levels. The comprehensive situation of the quality of English teaching: the principle of comprehensiveness is helpful to promote the comprehensive development of teachers’ English teaching level and the promotion of students’ quality education, but comprehensiveness does not include all relevant influencing factors into the English teaching quality evaluation system. All evaluation factors are scientifically and rationally screened, and the key influencing factors that affect the evaluation of students’ comprehensive quality are obtained, that is, the so-called evaluation indicators. While ensuring the comprehensiveness
of the English teaching quality evaluation system for teachers, it should have a certain direction, that is, the principle of pertinence. The English teaching quality evaluation system will train students to become talents with comprehensive quality of moral, intellectual, physical, and aesthetic development and meet the needs of society. The requirements of college graduates should pay more attention to the targeted cultivation of innovative practice spirit and ability so as to promote the development of college students. The quality evaluation of English teaching in colleges and universities is a scientific and reasonable means to stimulate the enthusiasm of teachers in English teaching and improve the quality of English teaching and finally achieve the purpose of improving the comprehensive quality of students. In the process of establishing the English teaching quality evaluation system, the differences of students should also be considered.

The same standard shall be adopted for all the objects participating in the evaluation, and the same number and level of evaluation individuals shall be determined for all the evaluation objects. In the quality education of colleges and universities, students are the main body of training and development. Only when teachers’ English teaching behavior and the knowledge content they teach are integrated into students’ cognitive structure and ability, students’ comprehensive quality can be truly improved. The design of evaluation indicators, the selection of methods, the way of collecting information, and the technologies and measures used in information processing must be objective and comprehensive, in line with reality, simple, and easy to implement, and save human, financial, and material resources.

This article combines information technology to construct a college English teaching system, improves the traditional English college teaching model, and improves the current English teaching effect.

2. Related Work

Higher education teaching is a complex system. Literature [4] pointed out that teaching quality can be roughly divided into four aspects: course content and structure, teaching time and methods, student learning and evaluation, and resource equipment. Literature [5] believes that college teachers’ teaching quality evaluation index system includes five aspects including teaching attitude, teaching content, teaching methods and means, teaching effect, and teaching status; literature [6] believes that the teaching quality evaluation index system of college teachers includes humanistic care, knowledge transfer, and ability training, three aspects of the content. Literature [7] believes that the teaching quality evaluation index system of college teachers includes humanistic care, knowledge transfer, and ability training, three aspects of the content. Literature [8] believes that the function of each constituent element in the system, and the overall effect of the interaction and mutual influence among the constituent elements. At present, the reform of teaching quality in colleges and universities is undergoing rapid changes in all directions: the training goal is from cultivating elites to cultivating citizens and workers [8]; the training concept is from imparting knowledge to cultivating ability and quality and then to the coordination of knowledge, ability, and quality. Development: the training mode has changed from narrow-caliber professional education to wide-caliber professional education and then to the combination of main and auxiliary, interdisciplinary electives, credit-based elective systems, etc.; classical liberal arts foundations to the frontiers of modern disciplines and majors, from closed knowledge systems to open time and space carriers, and to activities with a global perspective; leading, students “listen and practice test,” and then to teacher-led, student-centered, with the help of multimedia information, and teacher-student interaction, to help students to build knowledge, ability formation, and quality development process [9].

Literature [10] mentioned the research on PBL application teaching of VB programming and explored from theory and practice. However, there are very few evaluation studies on the teaching of PBL mode. Literature [11] aimed at the evaluation of the teaching effect of the problem-based learning teaching mode. The indirect inquiry survey method was used to participate in the questionnaire survey of the students of PBL teaching. Use PBL teaching effect to evaluate and analyze. There is little to no assessment and improvement of teaching quality through the PBL model. Literature [12] created and applied the PBL model in teaching and carried out a series of teaching model changes.

Literature [13] proposed that model construction requires the ability to visualize and process information. This shows that once the ability of students to reorganize information is improved, it is also necessary for model construction. Literature [14] conducted an empirical study on mental representation in the study of mental models and found that in the process of model construction, proper representation of it can help model construction, to solve problems, and cannot achieve the development of thinking. In addition, Borges pointed out that the ability to construct models is the mental model formed based on the natural phenomena and the representational ability extended under the mental model.

Literature [15] generalized the empirical evaluation tools FCI and MBT on the detection modeling effect. Literature [16] effectively helped students build models and improved their scientific inquiry ability. In literature [17], it is concluded that the student’s preconcept has nothing to do with his scientific thinking ability by means of a computer simulation system to build a model. Literature [18] constructs a variety of modeling environments and engages students in it, which promotes the transformation of students’ concepts and the ability of thinking development.
3. Research on Gain Control Algorithm of OFDM Receiver Based on English Informatization Teaching Needs

There are various nodes in the CRAHN network, and each node is moving randomly. Some nodes are very close to the receiver so that the received signal energy is very large, and the amplitude is too large, which may cause overload and distortion. Some nodes are far away from the receiver so that the received signal energy is very small, and the amplitude is too small, which may be ignored by the receiver as noise. In order to combat the near-far effect, gain control is required at the front end of the AD as shown in Figure 1.

Figure 2 is the AGC input and output transmission curve. When the input voltage amplitude is very small (AB section), the automatic gain control module does not work, and the gain is the maximum value. When the input voltage amplitude is within the controllable range (the BC section), the automatic gain control module takes effect to adjust the input signal through the automatic gain so that the output voltage amplitude changes within a small range. When the input voltage amplitude exceeds the range (CD section), the automatic gain control no longer takes effect, and the gain is the minimum value. Therefore, only when the input signal is within the controllable range, the automatic gain control module can take effect.

From the control signal point of view, automatic gain control can be divided into two types: digital and analog. Digital AGC control is flexible, but the noise performance is not good, and the receiver’s receiving sensitivity will be limited, so analog AGC is generally still used now. From the structure point of view, automatic gain control can be divided into two types: open-loop AGC and closed-loop AGC.

Automatic gain control can be divided into open-loop AGC and closed-loop AGC. Figure 3 is the open-loop AGC. Open-loop AGC is mainly composed of controlled amplifier/attenuator, threshold comparator, detector, and filter.

Figure 4 is the closed-loop AGC. Like open-loop AGC, closed-loop AGC is also composed of controlled amplifier/attenuator, threshold comparator, detector, and filter. Different from open-loop AGC, the input signal is first amplified or reduced by a controlled amplifier/attenuator, and then, the output signal is compared with a preset comparison voltage. Then, the output signal enters the detector for detection and finally enters the filter for filtering after detection.

Figure 5 is a digital automatic gain control circuit. In digital AGC, VGA is the core, and it affects the stability of the entire system and the response sensitivity of the input signal. Due to the relatively simple implementation, at present, the automatic gain control structure of the average absolute error controller is mainly used in the OFDM system. Improving the speed of convergence is the key to design and implementation.

Figure 6 is the automatic gain control based on the average absolute error structure, which is mainly composed of loop filter and power estimator. First of all, this paper uses the root mean square estimation method (equation (1)) to estimate the power of the data sampled in a fixed time [19].

$$P(n) = \frac{1}{W} \sum_{k=0}^{W-1} y^2(k).$$  (1)

In formula (1), $W$ is the sampling length and $y(k)$ is the sampling data.

$$G(n) = P_{\text{ref}} - P(n).$$  (2)

As shown in formula (2), the calculated power value $P(n)$ is subtracted from the reference power value $P_{\text{ref}}$, and the obtained difference $G(n)$ is sent to the loop filter as the gain value [20].

$$G(n) = K_1G(n) + K_2G(n-1).$$  (3)

In equation (3), $K_1$ and $K_2$ are the coefficients of the loop filter.
Figure 7 shows the amplitude-frequency characteristic curve of an ideal VGA. The bandwidth of the ideal VGA is only determined by the bandwidth of the input signal. Regardless of the gain value, the bandwidth of the variable gain amplifier is the same. At present, as the frequency of analog signals is getting larger and larger, even reaching the order of gigahertz, according to the characteristics of an ideal VGA, the bandwidth of the VGA must also reach the order of gigahertz. However, current technology often designs amplifiers with such a large bandwidth. Therefore, bandwidth expansion technology has been developed and popularized.

In the loop circuit of automatic gain control, the setup time of the loop needs to be kept constant. Therefore, it is necessary to ensure that the gain control of the variable gain amplifier is linear. However, the gain control range and control characteristics of an ideal VGA need to keep the slope \( K \) of the gain constant over the entire control voltage range as shown in Figure 8, that is,

\[
K = \frac{\text{Gain control range}}{\text{Control voltage range}} = \frac{G_{\max} - G_{\min}}{V_{c_{\max}} - V_{c_{\min}}} \quad (4)
\]

Figure 8: Gain control range and control characteristics of an ideal VGA.
Figure 10: Logarithmic domain model of automatic gain control loop.

$K$ is a constant, and the value of $K$ affects the performance of the system. As $K$ becomes larger, the loop time of automatic gain control becomes shorter, and as $K$ becomes smaller, the loop time of automatic gain control becomes longer. However, the value of $K$ depends on the actual circuit conditions.

In order to make the loop establishment circuit time of the automatic gain control independent of the input signal, the closed-loop automatic gain control of Figure 4 can be improved. In this paper, a logarithmic amplifier is added to the back of the detector and expressed in dB, which can make the system approximately linear. Figure 9 shows an automatic gain control loop with a logarithmic amplifier. The loop is a closed-loop circuit composed of five parts: variable gain amplifier, loop filter, comparison circuit, logarithmic amplifier, and detector.

$$V_{\text{out}} = G(v_c) \times V_{\text{in}},$$

(5)

where $V_{\text{in}}$ is the input signal voltage, $V_{\text{out}}$ is the output signal voltage, and $G(v_c)$ is the gain of the variable gain amplifier controlled by the voltage $v_c$. We assume that $K_{v1}$ is a constant with the same unit as $V_{\text{in}}$ and $V_{\text{out}}$ and transform equation (5) into logarithmic form as shown in the following equation:

$$\ln \left( \frac{V_{\text{out}}}{K_{v1}} \right) = \ln \left( G(v_c) \right) + \ln \left( \frac{V_{\text{in}}}{K_{v1}} \right).$$

(6)

This article uses formula (6) to change Figure 9 into a logarithmic block diagram. The logarithmic calculation formula is in the box.

In Figure 10, we can see

$$y(t) = 1\ln \left( G(v_c) \right) + x(t).$$

(7)

The transfer function of the loop filter $H(s)$ is

$$H(s) = \frac{GM}{sC}.$$  

(8)

According to Figure 10, the gain control voltage is

$$V_c = \int_0^t \frac{GM}{C} \left( V_{\text{ref}} - K_{v2} \right) dt.$$  

(9)

In this paper, the left and right sides of formula (7) are, respectively, derived, and formula (8) is inserted into it, where we can get

$$\frac{dy(t)}{dt} = \frac{dx(t)}{dt} + \frac{1}{G(V_c)} \frac{dG(V_c)}{dV_c} \frac{GM}{C} \left( V_{\text{ref}} - K_{v2} y(t) \right).$$

(10)

We turn the system into a linear system, as long as we put some restrictions on formula (10) and turn it into a linear equation, that is,

$$1 = \frac{G(V_c)}{dV_c} \frac{dG(V_c)}{dV_c} \frac{GM}{C} = K_x.$$  

(11)

The available time constant is

$$\tau = \frac{1}{K_x K_{v2}} = \frac{1}{1/G(V_c)(dG(V_c)/dV_c)(GM/C)K_{v2}}$$

(12)

Substituting formula (11) into formula (10), we can get

$$\frac{dy(t)}{dt} + K_x K_{v2} y(t) = \frac{dx(t)}{dt} + K_x V_{\text{ref}}.$$  

(13)

The settling time of automatic gain control remains constant. Therefore, in formula (12), $\tau$ is a constant, and we can get

$$\frac{1}{G(V_c)} \frac{dG(V_c)}{dV_c} = k_{g1} \Rightarrow d \ln(G(V_c)) = k_{g1} V_c.$$  

(14)

It can be seen from formula (14) that in the case of a logarithmic amplifier, to keep the automatic gain control setup time constant, the constraint is that there is an exponential relationship between the gain $G(V_c)$ of the VGA and the control voltage $V_c$. For systems with logarithmic amplifiers that satisfy equation (14), the settling time of automatic gain control gain is independent of the input signal, and the settling time meets equation (12).

Using a logarithmic amplifier, that is, there is an exponential relationship (dB-linear relationship) between the gain $G(V_c)$ of the variable gain amplifier and the control voltage $V_c$. Regardless of the gain value, the setup time of the automatic gain control gain can be made independent of the input signal, and it is a constant. Moreover, the dynamic range of automatic gain control is greatly improved.
In Ad602, the input signal $V_{in}$ ($V_{GHI} - V_{GIL0}$) is controlled by the AGC control voltage $V_{Gain}$ to control the attenuation amplitude, and after being attenuated by the R-2R ladder network, the output voltage $V_{out}$ can be obtained as shown in the following equation:

$$\text{Gain} = 32 \times V_{gain} + 10 = 32 \times (V_{ALH1} - V_{ALL1}) + 10. \quad (15)$$

A resistor with $R = 40.2k\Omega$ is connected between pin4 and pin5 of AD8310 so that the slope of the AD8310 curve is $3.125mV/dB$. When the input signal $V_{in}$ unit is dBm and the output signal $V_{out}$ unit is mV, the following formula can be obtained:

$$V_{out} = 31.25mV/dB \times (V_{in} + 95dBm). \quad (16)$$

The voltage conversion into logarithm and logarithm conversion into voltage can be designed as a table. The internal resistance of the chip is fixed, so you can use the look-up table to determine the corresponding relationship, that is, this paper converts $V_{in}$ (unit: V) into $x(t)$ (unit: dBm) and $y(t)$ (unit: dBm) into $V_{out}$ (unit: V).

According to Figure 11, we can get

$$\begin{align*}
V_{out} &= G(v_c) \times V_{in}, \\
y(t) &= \ln(G(v_c)) + x(t) = \text{Gain1} + \text{Gain2} + x(t) = \text{Gain} + x(t).
\end{align*} \quad (17)$$

According to formula (15), we can get

$$\begin{align*}
\text{Gain} &= -10 \text{ dB} \quad (V_{\text{gain}} < -0.625 \text{ V}), \\
\text{Gain} &= (-10 \text{ dB}, 30 \text{ dB}) \quad (-0.625 \text{ V} \leq V_{\text{gain}} \leq 0.625 \text{ V}), \\
\text{Gain} &= 30 \text{ dB} \quad (V_{\text{gain}} > 0.625 \text{ V}).
\end{align*} \quad (18)$$

When the output $V_{g}$ of AD8310 is 0.975 V ~ 2.225 V, then $V_{ALL1}$ and $V_{ALH1}$ are 0.975 V ~ 2.225 V. According to the formula $V_{\text{gain}} = V_{ALH1} - V_{ALL1}$, the control voltage $V_{\text{gain}}$ of the first stage of AD602 is 0.625 V ~ 0.625 V, and the gain is 30 dB ~ -10 dB. The control voltage $V_{\text{gain}}$ of the second stage of AD602 is 1.875 V ~ 0.625 V, the gain is 30 dB unchanged, and the total gain of the two stages is 60 dB ~ 20 dB.

When the output $V_{g}$ of AD8310 is 2.225 V ~ 3.475 V, then $V_{ALL2}$ and $V_{ALH2}$ are 2.225 V ~ 3.475 V. According to the formula $V_{\text{gain}} = V_{ALH1} - V_{ALL1}$, the control voltage $V_{\text{gain}}$ of the first stage of AD602 is -0.625 V ~ -1.875 V, and the gain is -10 dB unchanged. The control voltage $V_{\text{gain}}$ of the second stage of AD602 is 0.625 V ~ 0.625 V, the gain is 30 dB ~ -10 dB, and the total gain of the two stages is 20 dB ~ -20 dB.

In summary, when the output $V_{g}$ of AD8310 is 0.975 V ~ 3.475 V, the gain of AD602 is 60 dB ~ -20 dB, the dynamic range is 80 dB, and the gain $G_{ALL}$ and $V_{g}$ meet the linear relationship as shown in the following formula:

$$G_{ALL} = -32 \times V_{g} + 91.2 \text{ dBm}. \quad (19)$$

Next, we will discuss whether it is possible to provide an appropriate gain GAIN for the input voltage $V_{in}$ within a reasonable range so that the design can obtain a stable output value $V_{out}$. We assume that the system needs to maintain a constant output voltage $V_{out}$ of 500 mV (the logarithm is expressed as -3.8 dBm by looking up the table), and the gain of the system is -20 dB ~ 60 dB. Then, according to formula (15), the available input voltage range must meet -63.8 dBm ~ 16.2 dBm, AD602 can provide suitable gain value GAIN.

When the input voltage value of $V_{in}$, its logarithm is expressed as $x(t)$dBm, and formula (16) is inserted into formula (17) to obtain

$$G_{ALL} = -32 \times V_{g} + 91.2 \text{ dBm},$$

$$= -32 \times [31.25mV/dB \times (x(t) + 95\text{dBm})] + 91.2\text{dBm},$$

$$= -x(t) - 3.8\text{dBm}. \quad (20)$$

That is, $G_{ALL} + x(t) = -3.8 \text{ dBm}$ is established, and the desired gain is the difference between the output level and the input level. Local autocorrelation algorithm is shown in Figure 12.

$$m(n) = \sum_{k=0}^{L-1} r^*(n-k) \times \text{LTS}(k). \quad (21)$$

The relative energy value is $m(n)$, and we assume that the reference threshold is $Th$. When $m(n) > Th$, it is judged that there is a frame. The peak point of $m(n)$ is the starting point of the signaling domain.

This article implements the formula according to IDFT as follows:

$$x(n) = \sum_{k=1}^{N} X(k)e^{j2\pi km/N}. \quad (22)$$

When the even-numbered subcarriers send data, we get...
The measurement function is used to achieve synchronization, and the synchronization signal from the complex signal training sequence random weighting factor to form a new synchronization domain PN sequence, the transmitter introduces a pseudo-location of the sampling point.

At the receiving end, the delay-related ML algorithm is used to achieve synchronization, and the synchronization measurement function is

$$M(n) = \frac{|C(n)|^2}{P(n)}$$  \hspace{1cm} (25)

Among them, there are

$$C(n) = \sum_{m=0}^{L-1} y_{n+m}^* y_{n+m+L},$$  \hspace{1cm} (26)

$$P(n) = \sum_{m=0}^{L-1} |y_{n+m}|^2.$$  \hspace{1cm} (27)

In the SC synchronization algorithm based on the time-domain PN sequence, the transmitter introduces a pseudo-random weighting factor to form a new synchronization training sequence $$T = [A_i B_i]$$, where $$B_i$$ and $$A_i$$ are of length $$L$$, and both satisfy $$B_i(i) = A_i(i) \times PN_i, i = 0, 1, \ldots, L-1$$, and $$L$$ is the subcarrier Number, $$PN_i$$ is a pseudo-random sequence of length $$L$$, with a value of +1 or −1.

At the receiving end, we extract the synchronization signal from the complex signal $$y_n$$ after quadrature demodulation as shown in the following formula:

$$C(n) = \sum_{m=0}^{L-1} PN_m Z(n+m),$$  \hspace{1cm} (28)

Among them, the value of the correlation function $$C(n)$$ is the correlation value of the previous sequence multiplied by the PN sequence and the next sequence, and $$N$$ is the location of the sampling point.

We assume $$Z(n+m) = y_{n+m}^* y_{n+m+L}$$, that is, the conjugate value of $$L$$ points apart, then formula (28) can be changed to

$$C(n) = \sum_{m=0}^{L-1} PN_m Z(n+m).$$  \hspace{1cm} (29)

It can be seen from formula (29) that compared with the SC algorithm, the receiving end of the SC algorithm based on the time-domain PN sequence multiplies the conjugate sum of length $$L$$ by the PN sequence of length $$L$$ and then takes the sum. The correlation function $$C(n)$$ is derived.

In order to eliminate the interference of the signal amplitude at the receiving end to the synchronization signal, the above equation is processed as follows:

$$M(n) = \frac{|C(n)|^2}{P^2(n)}$$  \hspace{1cm} (30)

Among them, there are

$$P(n) = \sum_{m=0}^{L-1} |y_{n+m}|^2.$$  \hspace{1cm} (31)

In the formula, $$P(n)$$ is the energy sum of these $$L$$ points, and the synchronization point is estimated as

$$\tilde{\epsilon} = \arg\max_n(M(n)).$$  \hspace{1cm} (32)

The SC synchronization algorithm based on the time-domain PN sequence can get the precise synchronization point. However, in the broadband OFDM transmission system, the value of $$L$$ is very large, and the amount of calculation required to implement the SC synchronization algorithm based on the time-domain PN sequence is also very large, which is often impossible to implement in hardware. At the receiving end, every time a set of orthogonal demodulation values is obtained, the value of $$M(n)$$ needs to be calculated once, that is, the values of $$C(n)$$ and $$P(n)$$ are calculated once. However, it takes a huge amount of calculation to calculate $$C(n)$$ once.

$$C(n) = \sum_{m=0}^{L-1} PN_m Z(n+m),$$

$$= PN_0 Z(n) + PN_1 Z(n+1) + \cdots + PN_{L-1} Z(n+L-1),$$  \hspace{1cm} (33)

$$C(n+1) = \sum_{m=0}^{L-1} PN_m Z(n+1+m),$$

$$= PN_0 Z(n+1) + PN_1 Z(n+2) + \cdots + PN_{L-1} Z(n+L).$$  \hspace{1cm} (34)

The value of the PN sequence is fixed at +1 or −1, and the value of $$Z(n+m)$$ changes with the change in sampling point.
position. It can be seen from (33) and (34) that the value of $C(n + 1)$ cannot be obtained by the value of $C(n)$. However, it needs to read and subtract the value of $L$ $Z(n + m)$ for the PN sequence. This can be seen that the amount of calculation is huge and increases with the increase of $L$.

The SC algorithm based on the time-domain PN sequence has a large amount of calculation and takes up too much FPGA hardware resources. This paper improves the SC algorithm based on the time-domain PN sequence. In the SC algorithm, there are

$$C(n) = \sum_{m=0}^{L-1} y_{n+m}^* y_{n+L+m}.$$  

(35)

We assume $Z(n + m) = y_{n+m}^* y_{n+L+m}$, then there are

$$C(n) = \sum_{m=0}^{L-1} Z(n + m).$$  

(36)

We can get

$$C(n + 1) = C(n) + Z(n + L) - Z(n).$$  

(37)

It can be seen that the SC algorithm is simple. This is because the $C(n)$ value of each sampling point can be obtained by adding or subtracting two conjugate values from the previous sampling point. Combining the SC algorithm and the SC algorithm based on the time-domain PN sequence, this paper divides the received $L$ data into $P$ groups, and each group of $K$ data is multiplied by the same PN value. This article regards it as dividing it into $P$ groups of length $K$ to calculate the $C(n)$ value of each segment using the SC algorithm and adding the $C(n)$ values of each group to obtain the $C(n)$ value of $L$ data. The formula (35) is transformed into

$$C(n) = \sum_{i=0}^{p-1} PN_i \sum_{m=0}^{k-1} y_{n+m+ik}^* y_{n+L+m+ik}.$$  

(38)

We assume $C_i(n) = \sum_{m=0}^{k-1} y_{n+m+ik}^* y_{n+L+m+ik} = \sum_{m=0}^{k-1} Z(n + ik + m)$, and then, formula (38) can be changed to

$$C(n) = \sum_{i=0}^{p-1} PN_i C_i(n).$$  

(39)

Among them, $p$ is the length of the PN sequence, $L$ subcarriers are divided into $P$ segments evenly by the PN sequence, and the number of subcarriers in each segment is $k$ ($L$, $p$, and $k$ are all integers). We assume that the frequency deviation at both ends of the transceiver is $\Delta f_v$, and the phase deviation is $\theta_0$. In this way, there will be a phase deviation of $e^{j(\Delta f_v \tau)}$ in the intermediate frequency output signal of the receiving end. The influence of this deviation on the phase is cumulative and will become larger over time. The carrier synchronization module can estimate $\Delta f_v$ and perform frequency deviation correction. The channel estimation and equalization module can estimate $\theta_0$.

We assume that the transmitted signal is $a_n + jb_n$, after IFFT transformation, it is

$$a_n + jb_n = \frac{1}{N} \sum_{n=0}^{N-1} (a_n + jb_n) e^{j(2\pi kn/N)}.$$  

(40)

Because $N = (1/T_s \cdot \Delta f)$, $t = k \cdot T_s$, $\Delta f = 2\pi \Delta f$, equation (40) can be reduced to the following equation:

$$a_n + jb_n = \frac{1}{N} \sum_{n=0}^{N-1} (a_n + jb_n) e^{j\Delta \omega t}.$$  

(41)

In general cable channels and slow-moving rural channels, the channel conditions are better, and only Gaussian white noise and the effects of multipath and Doppler shift are not considered. After quadrature modulation, Gaussian white channel, and quadrature demodulation, equation (41) is transformed into the following equation:

$$\left( \frac{1}{N} \sum_{n=0}^{N-1} (a_n + jb_n) e^{j\Delta \omega t} \cdot e^{-jw_t^t + n(t)} \right) e^{j(w_t + \Delta w_t + \theta_0)}$$

$$= \frac{1}{N} \sum_{n=0}^{N-1} (a_n + jb_n) e^{j\Delta \omega t + \Delta w_t + \theta_0} + n(t) e^{j\left(w_t + \Delta w_t + \theta_0\right)}.$$  

(42)

After low-pass filtering, equation (42) is transformed into the following equation:

$$\frac{1}{N} \sum_{n=0}^{N-1} (a_n + jb_n) e^{j\Delta \omega t + \Delta w_t + \theta_0} - \frac{1}{N} \sum_{n=0}^{N-1} (a_n + jb_n) e^{j\Delta \omega t} \cos(\Delta w_t + \theta_0)$$

$$+ j \sin(\Delta w_t + \theta_0)).$$  

(43)

Due to the existence of $e^{j(\Delta w_t + \theta_0)}$, the signal entering the FFT module is not a quadrature signal. Therefore, when the frequency deviation is large, the influence of the frequency deviation cannot be ignored. It can be seen from equation (43) that the frequency offset is inversely proportional to the duration. When the frequency deviation is large, the frame length can be reduced to reduce the influence of the frequency deviation.

In this paper, in the local autocorrelation scheme, the $I$ signal and the $Q$ signal after the conjugate are multiplied by the square sum operation can eliminate the influence of the frequency offset under the Gaussian channel.

In this paper, formula (43) is written as

$$(I(t) + jQ(t)) \cdot \left( \cos(\Delta w_t + \theta_0) + j \sin(\Delta w_t + \theta_0) \right)$$

$$= I(t) \cos(\Delta w_t + \theta_0) - Q(t) \sin(\Delta w_t + \theta_0)$$

$$+ j[I(t) \sin(\Delta w_t + \theta_0) + Q(t) \cos(\Delta w_t + \theta_0)].$$  

(44)

This article multiplies this signal with the local training sequence $I_k(T) + jQ_k(T)$ conjugate, and we assume that $M = I(t) \sin(\Delta w_t + \theta_0) + Q(t) \cos(\Delta w_t + \theta_0), N = I(t) \cos(\Delta w_t + \theta_0) - Q(t) \sin(\Delta w_t + \theta_0)$, and we get

The $I$ signal is
\[ r_i(t) = Q_L(t) \times M + I_L(t) \times N. \] (45)

The Q signal is
\[ r_q(t) = I_L(t) \times M - Q_L(t) \times N. \] (46)

Adding the square of I road to the square of Q road, we get
\[ r^2(t) = r_i^2(t) + r_q^2(t), \]
\[ = \left[ I_L^2(t) + Q_L^2(t) \right] \cdot \left[ I^2(t) + \sin^2(\Delta w_r t + \theta_0) + \cos^2(\Delta w_r t + \theta_0) \right], \]
\[ = \left[ I_L^2(t) + Q_L^2(t) \right] \cdot \left[ I^2(t) + Q_r^2(t) \right]. \] (47)

It can be seen from equation (47) that taking the sum of the squares of the I and Q channels can completely cancel the influence of frequency offset, and the low-pass filter after quadrature demodulation at the receiving end can filter out the influence of Gaussian white noise on the local auto-correlation. In the SC algorithm, because the conjugate correlation method is adopted, the deviation can also be eliminated.

After the multipath channel, the formula (47) is transformed into
\[ A_1 N \sum_{n=0}^{N-1} (a_n + j b_n) e^{j \Delta u + \Delta w_r t + \theta_n}, \]
\[ + \frac{A_2}{N} \sum_{n=0}^{N-1} (a_n + j b_n) e^{j \Delta u + \Delta w_r t + \theta_2} + \cdots \]
\[ + \frac{A_m}{N} \sum_{n=0}^{N-1} (a_n + j b_n) e^{j \Delta u + \Delta w_r t + \theta_m}. \] (48)

Due to a series of influences such as the difference between the receiving and sending crystals and the Doppler frequency offset, \( \Delta u \) is different, the value is very large, its value cannot be ignored, and with the increase of time, the influence of the phase deviation becomes larger and larger. \( \theta_n \) is also different, and it is not easy to estimate. Therefore, the influence of frequency offset in a multipath Doppler channel is much more complicated than that in a Gaussian channel.

On the basis of distributed structure reorganization and phase space reconstruction preprocessing of high-dimensional data information flow, the optimization design of fuzzy information extraction algorithm for big data is carried out, and the expression of Lorenz chaotic attractor is given in the following equation:
\[
\begin{align*}
\frac{dx}{dt} &= -\sigma x + \sigma y, \\
\frac{dy}{dt} &= -xz + rx - y, \\
\frac{dz}{dt} &= xy -bz.
\end{align*}
\] (49)

Among them, each quantity is a reduced dimensionless quantity, \( t \) represents the sampling time of the fuzzy data; \( x, y, z \) represent the partition variables of the Lorenz chaotic attractor; and \( \sigma, r, b \) are the fuzzy constraint parameters. The fourth-order Runge–Kutta method is used for chaotic partitioning of fuzzy information. For the feature vector \( X_n \) of fuzzy information, it selects the nearest neighbor trajectory vector \( X_{n(n)} \) and obtains the Euclidean distance of the cluster center area of the chaotic partition of the big data as
\[
R_{mn} = \| X_{n(n)} - X_{n(n)}^{(m)} \|_2 = \min_{j=0;N_n-1, n \neq n} X_n - X_{n}^{(m)}_n.
\] (50)

Adaptive learning is carried out in the above reconstructed phase space, and when the embedding dimension of the phase space is increased from \( m \) to \( m+1 \), the optimal value of the chaotic partition cluster center for fuzzy information extraction is obtained, which is expressed as
\[
R_{(m+1)n} = \| X_{n(n)} - X_{n(n)}^{(m+1)} \|_2 = \frac{m-1}{\sum_{l=0}^{m-1} \left( X_{n(n)+l} - X_{n+l} \right)^2}.
\] (51)

If \( R_{(m+1)n} \) is much larger than \( R_{mn} \), the interclass attribute disturbance of fuzzy information does not affect each other, and it can be used as the fuzzy clustering center. According to this characteristic, the obtained \( X_{n(n)} \) is the nearest neighbor of \( X_n \). In this paper, the decision statistics are constructed. In order to ensure the accuracy of the decision, the threshold judgment for the extraction of modular amide information is carried out in the chaotic division, and the following decision criteria are obtained:

- **Criterion 1** is
  \[
  \sqrt{\frac{R_{(m+1)n}^2 - R_{mn}^2}{R_{(m+1)n}^2}} = \frac{R_{mn}}{R_{(m+1)n}} \geq R_{tol}.
  \]
- **Criterion 2** is
  \[
  \sqrt{\frac{(1/N) \sum_{k=1}^{N} x_k^2}{(1/N) \sum_{k=1}^{N} x_k^2}} > A_{bol}.
  \]

Among them, \( R_{tol} \) is the partition threshold of the Lorenz chaotic attractor. According to the empirical value, \( R_{tol} \) can take 15; \( A_{bol} \) represents the judgment threshold of fuzzy information chaotic partition, and it takes \( A_{bol} = 2 \).

According to the above analysis, the chaotic partition of fuzzy information in the big data environment is realized, and the adaptive learning training of fuzzy information...
extraction from big data is carried out by combining decision criteria and decision statistics, and feature extraction and information extraction are carried out.

We assume that the distribution time series \( \{X_n\} (n = 1, 2, \ldots, N) \) of fuzzy information represents the original large data feature distribution set to be partitioned. Under the chaotic partition processing, the characteristic distribution after the chaotic partition is obtained \( X_n = X_n + \eta \), where \( \eta \) is the observation noise. In the distributed information sources of \( d \) big data, the phase space reconstruction technology is used to perform autocorrelation feature matching on the chaotic attractor load feature quantities of the big data, and the feature matching output is obtained.

\[
X_n = \{X_n, X_{n-1}, X_{n-2}, \ldots, X_{n-(d-1)}\}. \tag{52}
\]

We set \( R_{d,L} \) to be the matrix of \( d \times L \), and the average mutual information feature quantity of fuzzy information extracted in the closed frequent term region of the chaotic partition is

\[
R_1 = [X_1, X_2, X_3, \ldots, X_d]^T. \tag{53}
\]

The association rule vector set of fuzzy information is

\[
R_1^T R_1 = [X_1, X_2, \ldots, X_m][X_1, X_2, \ldots, X_m]^T. \tag{54}
\]

Using the singular decomposition method to decompose the eigenvalues of fuzzy information, we get

\[
R_1^T R_1 = V_1 \Sigma_1 V_1^T. \tag{55}
\]

For \( L + 1 \) to \( 2L \)-dimensional chaotic partitioned big data, according to the above method analogy, the output characteristic value of fuzzy information extraction is obtained

\[
R_1^T R_2 = V_2 \Sigma_2 V_2^T, \tag{56}
\]

\[
R_2 = [X_{d+1}, X_{d+2}, \ldots, X_{d+m}]^T, \tag{57}
\]

\[
R_2^T R_2 = [X_{d+1}, X_{d+2}, \ldots, X_{d+m}][X_{d+1}, X_{d+2}, \ldots, X_{d+m}]^T. \tag{58}
\]

Among them, the test set of fuzzy information

\[
V = [V_1, V_2, \ldots, V_m] \in \mathbb{R}^{mm} \text{ is orthogonal, that is, } V V^T = I_m; \Sigma = \text{diag} (\sigma_1, \sigma_2, \ldots, \sigma_m) \in \mathbb{R}^{mm}. \]

The fuzzy eigenvector set \( R^T R \) of the association rules after the chaotic partition satisfies the balance between classes, and the cluster fusion degree of the extracted eigenvalues is sorted as follows:

\[
\sigma_1 > \sigma_2 > \sigma_3 > \cdots > \sigma_{s_1} > \sigma_m. \tag{59}
\]

It can be seen from the analysis that the dimension of a certain information extracted by the above method is high, and feature compression is required. The steps of feature compression are described as follows:

(a) The algorithm calculates the \( l \)-dimensional fuzzy information feature vector \( X(l), n_i \) in the chaotic partition area. According to the clustering attribute value of the partition, the fuzzy information

\[
\tilde{w}_i = \frac{1}{n_i} \sum_{k=1}^{n_i} \left( X_k^{(l)} - m \right) \left( X_k^{(l)} - m \right)^T. \tag{60}
\]

(b) According to the center vector of the chaotic partition attractor, the discrete degree matrix \( S_b = \sum_{i=1}^{c} p_i (m - m)^T \) of fuzzy clustering is obtained, and the weighted average of \( m = \sum_{i=1}^{c} p_i m_i \) is obtained.

(c) The algorithm adopts the decision statistical regression analysis method, performs feature compression in the high-dimensional space, and outputs the feature quantity \( J(X_i) = (y_i^T S_y y_i / \lambda_i) (j = 1, 2, \ldots, l) \) in the training set, outputs \( y_j (j = 1, 2, \ldots, d) \), and obtains the fuzzy transformation weighting matrix \( W = [y_1, y_2, \ldots, y_d] \).

(e) The information extraction result of the algorithm output feature compression is \( \bar{X} = W^T X \).

Through the above processing, the feature quantity of fuzzy English information output after the chaotic partition is reduced from \( l \) to \( d \) dimension, thus reducing the computational cost.

4. College English Intelligent Teaching System

The intelligent teaching system is established on the basis of effective organization of the learning objectives, learning resources, learning objects, and student characteristics in different teaching processes and links. Teachers use the teacher interface to guide, suggest, and evaluate to keep abreast of students’ course progress and cognitive level and give students maximum help and feedback. This constitutes the various modules and relationships of ITS as shown in Figure 13(a).

The model describes various attribute mapping entities in the knowledge base, including address descriptions of various learning resources, learning goals, learning objects, and context characteristics that match the student model and teacher model. In addition, it includes attributes such as resource acquisition channels, teaching activity sequences, and knowledge presentation strategies that match the teacher model. The specific description is shown in Figure 13(b).

After constructing the above model, we will calculate the wireless reception effect of English information, the digitization effect of English teaching resources, and the teaching effect of this model, and the results are shown in Figures 14–16.
Figure 13: Structure of intelligent English teaching system. (a) System module structure and (b) knowledge base storage structure.

Figure 14: The effect of wireless reception of English information.

Figure 15: Digital effect of English teaching resources.
Through the above experimental research, it can be known that the college English teaching system based on information technology proposed in this paper has a good teaching effect.

5. Conclusion

The purpose of the evaluation of college English teaching quality is to arouse the enthusiasm of teachers and improve their own English teaching level and English teaching quality. In the process of evaluating the quality of college English teaching, certain principles should be followed. Moreover, the key role of evaluation principles for the effectiveness and reliability of evaluation results is the basic requirement for evaluation. The objectivity of the evaluation results of English teaching quality is the key to the evaluation of English teaching quality. At the same time, it is necessary to collect evaluation data in a true and comprehensive manner; conduct evaluations objectively, impartially, and truthfully, and avoid personal feelings in the evaluation process. This article combines the information technology to construct the college English teaching system and improve the traditional English college teaching mode and the current English teaching effect. The experimental research results show that the college English teaching system based on information technology proposed in this paper has a good teaching effect.

Data Availability

The labeled data set used to support the findings of this study is available from the corresponding author upon request.

Conflicts of Interest

The authors declare no conflicts of interest.

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