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Broadband signal generator for the approximation of a magnetotelluric source for indoor testing

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Abstract
To test the frequency response of a magnetotelluric (MT) receiver, a broadband source, especially white noise is more efficient and intuitive than single frequency signals. In view of the absence of an appropriate source generator for MT receiver indoor testing, we designed a broadband signal generator based on a pseudo-random binary sequence (PRBS). Firstly, we divided the whole MT band into two segments to avoid data redundancy and simplify calculation in data processing and designed a generator composed of several modules: a clock module, a PRBS logic module, and a voltage level conversion module. We conducted a detailed analysis of the optimal parameter selection methods for each module, and key parameters including clock frequency, order, the primitive polynomial and the original states of the linear registers were determined. The generator provides four-channel PRBS signals with two effective bandwidths of $5 \times 10^{-3}$–714 Hz and 0.1 Hz–14 kHz which are broad enough to cover the frequency range for different MT methods. These four-channel signals were used to simulate two modes of sources ($xy$ and $yx$) with strong auto-correlation and weak cross-correlation. The power spectral density is quite stable in the whole passband. The new generator is characterized by broadband output in low-frequency bands, low power consumption, simple operation and reliable performance. Indoor and field tests indicated that the generator can provide an analog MT source and is a practical tool for MT receiver indoor testing.

Keywords: magnetotelluric sounding, PRBS, parameter selection, circuit design

1. Introduction
The magnetotelluric (MT) method operates in a very broad frequency range and has played an important role in geophysical prospecting (Fuji-ta et al 1999, Deng et al 2003, 2013). It is an indispensable means of deep geophysical exploration and has been widely used in geothermal field investigation as well as earthquake forecasting and prediction (Tank et al 2005, Monteiro Santos et al 2006, Zhang et al 2014).

After several decades of development, there are three categories of MT sounding depending on their different source or frequency band: natural-source magnetotelluric (NSMT) utilizing a broad frequency band of $1 \times 10^{-3}$–$1 \times 10^3$ Hz (Zhang and Pedersen 1991), audio-frequency magnetotelluric (AMT) proposed by Berdichevsky in the 1960s (Berdichevsky 1969) with a bandwidth of 1–10 kHz, and controlled-source audio magnetotelluric (CSAMT) proposed by Strangway and Goldstein in the 1970s (Goldstein and Strangway 1975) which uses the same measuring method as the AMT method but has a greater strong anti-interference capacity (Bastani et al 2011, Younis et al 2015).

The successful application of MT sounding is dependent on the acquisition of high-quality data. In recent years, with the rising difficulty of resource exploration, MT instruments have been constantly upgraded to meet the requirements of searching for natural resources. The resolving ability of MT
methods depends on the frequency band of the measurements (Key and Constable 2002), but the system responses vary at different frequencies and the frequency responses of different channels are never exactly the same. The sensitivity and frequency characteristics of the magnetic sensors also differ for different frequencies. To cover the whole frequency band of the natural electromagnetic field while keeping high sensitivity and high signal-to-noise ratio (SNR), MT receivers tend to be equipped with a low-noise operational amplifier at the high-frequency band ($1 \times 10^2$–$1 \times 10^4$ Hz). However, the $1/f$ noise of the pre-amplifier appears in the low-frequency band ($<100$ Hz), and the lower the frequency the higher the power spectrum of this noise. As a result, the measurement accuracy will be discounted. In addition, to ensure the reliability of the measuring results, periodic calibration should be conducted during the long-term use of the receivers. Thus, any new MT receiver should be tested before actual field operation to establish these profiles.

In addition to routine tests for equivalent input noise, dynamic range, pulse test, amongst others, testing of the frequency response is also required to verify the performance of the new receiver and to ensure that the instrument’s frequency band is consistent with that of the measured signal for different MT observations. During the testing process, an appropriate input is very important for the validity of the testing. Thus, knowledge of system performance in the presence of an excitation signal is important for characterizing the precision and accuracy of MT receivers. Therefore, we designed a signal generator to translate this conceptualization into a real signal for indoor testing. The new generator assisted the measurement of the frequency response for a newly developed MT receiver in a frequency band covering several decades. Consequently, the lossless operating frequency band and the frequency response characteristic can be determined after indoor testing.

Previously, a frequency sweeping method using single frequency signals, sine or square waves, has been used to test the amplitude and phase frequency response of MT receivers. However, this approach is very inefficient and time-consuming in relation to covering the whole bandwidth. This linearity of the frequency sweep usually decreases with an increasing sweep bandwidth, and the sweep nonlinearities may increase the testing variance. Furthermore, it is difficult to obtain all of the desired frequencies across the whole MT band using one signal generator. A broadband source would be a better choice for testing over the full bandwidth. We need a source to simulate the MT field, which can quantify the MT receiver’s working performance. After indoor testing, a new receiver is able to be used for fieldwork.

The frequency spectrum of the MT receiver’s target—the MT field—shows the uniform distribution in a very broad band, and has the characteristic of white noise. White noise has sharp auto-correlation, weak cross-correlation, and its power spectrum is uniform across the frequency band. This means it is an ideal excitation source for MT receiver testing, and could test all the frequencies across a wide frequency range at the same time, making the test more convenient and accurate. However, it is very difficult to obtain reproducible white noise with high performance and the appropriate source device is also complex. Rader et al. (1970) put up with a quick method to generate white noise by using the exclusive-OR gate and realized the digital hardware design for this generator. But, the generators only output short-period noise and do not produce multichannel independent noise. With the development of digital computation, more and more researchers attempted to generate white noise by using digital technology. At the same time, the development of microchips provided a basic foundation for the hardware realization of digital white noise generators. Signals generated by this method, called a ‘pseudo-random sequence’ (PRS), have similar spectral properties as true random white noise. A PRS is a repeatable signal and very suitable for use as a stimulus source for MT receiver indoor testing. Thus, we chose a PRS controller to produce matched signals with similar statistical characteristics to band-limited white noise. In practice, a PRS is generated using a binary sequence $\{0, 1\}$, known as a pseudo-random binary sequence (PRS). A PRBS is a deterministic signal with properties determined by clock frequency $f_c$, and the length of the shift register (order $n$) (Amrani et al. 1998). Many researchers attempted PRBS generation. Smith and Hamilton (1966) designed a digital pseudo-random noise generator using a linear-feedback shift register and exclusive-OR (modulo-two) logic elements. Neuvo and Ku (1975) designed a noise generator to produce Gaussian background noise by first generating a basic PRBS and then adding samples from the PRBS generator. Mahmood and Abdul-Sada (1987) built a programmable noise generator by using an EPROM (electronically programmable read only memory) memory and made it possible to change the probability density function without any major change in the required hardware. These early instruments suffered a general lack of high performance due to technical limitation. Sliwczynski (2007) designed a PRBS generator running at 1.5 Gbps using the high-speed serializer TLK2201B. Bouvier et al. (2014) developed a PRBS generator operating at a speed close to 100 Gbit s$^{-1}$ based on 290 GHz-FT InP HBT technology. However, the effective frequency bands of these modern PRBS generators mainly distributed in very high-frequency domains ($>1 \times 10^6$ Hz), and the low-frequency band was insufficient. Their bandwidths did not match with that of the MT field, especially low-frequency components and they are not suitable for MT receiver testing. Therefore, we designed a broadband signal generator specifically for indoor MT field simulation.

To simulate the four-channel collected MT signals, we designed a four-channel PRBS generator which provides broadband signals (independently for mode $xy$ and mode $yx$) with two alternative bandwidths $5 \times 10^4$–$5 \times 10^7$ Hz and $0.1$–$1 \times 10^2$ Hz (see figure 1). We determined the key indexes of the required PRBS based on a theoretical analysis, and designed the hardware for the new generator on this basis. Parameter selection methods, hardware design, and the subsequent generator design are then described in detail. Indoor and field tests show that the power spectral density of the signals produced by this generator has the characteristics of white
noise and was applicable to MT receiver testing. Finally, we summarize our outcomes and provide conclusions.

2. Method

2.1. PRBS theory introduction

PRBS signals can imitate white noise as \( n \) increases. They may also be constructed to cover a broad frequency band that can be adjusted by sequence size and \( f_c \). Common PRBSs include the m-sequence, M-sequence and gold-sequence. The m-sequence is the most basic and important PRS, and the last two are derived from this sequence by calculation. The m-sequence has more desirable auto-correlation and cross-correlation than the gold-sequence, and the device for the m-sequence has a lower hardware cost compared to the M-sequence. Hence, we used the m-sequence for our signal generator.

The m-sequence, also called ‘maximum length sequence’, is produced by maximal linear-feedback shift registers and has valid lengths of \( N = 2^n - 1 \). The auto-correlation function \( R_{cc} \) of an m-sequence with amplitude \( A \), clock frequency \( f_c \), and register length \( n \) is governed by equation (1):

\[
R_{cc}(\tau) = \begin{cases} 
  A^2 \left[ 1 - \left( \frac{1}{N} \right) \times \tau \times f_c \right] & 0 \leq \tau \leq \frac{1}{f_c} \\
  -\frac{A^2}{N} & 1/f_c \leq \tau \leq (N-1)/f_c \\
  A^2 \left[ \left( 1 - \frac{1}{N} \right) \times \tau \times f_c - N \right] & (N-1)/f_c \leq \tau \leq N/f_c 
\end{cases}
\]

where, \( \tau \) is the time-shifting.

Let \( T = N/f_c \) and \( f_0 = f_c/N \). Then, the corresponding PSD \( P_i \) is

\[
P_i = \frac{1}{2T} \int_{-T}^{T} R_{cc}(\tau) e^{-2\pi i f_0 \tau} d\tau
= \frac{1}{T} \int_{0}^{T} R_{cc}(\tau) \cos 2\pi f_0 \tau \, d\tau, (i = 1, 2, \ldots N) \quad (2)
\]

And if we set \( \Delta t = 1/f_c \) and \( \omega_0 = 2\pi f_0 \), then:

\[
P_i = \frac{A^2}{\Delta t} \left( \frac{1}{N} \right) \cos \omega_0 \tau \, d\tau
= \frac{1}{N} \int_{0}^{\Delta t} \sin \left( \frac{\pi n f_0}{f_c} \right)^2 (i = 1, 2, \ldots) \]

The PSD amplitude envelope is \((\sin \pi f_0/f_c)^2/\pi f_0/f_c\)^2, indicating that the power at each frequency point is inversely proportional to \( n \).

The PSD of an m-sequence is shown in figure 2. The PSD is a line spectrum with line spacing of \( 1/2^n - 1 \), composed of the fundamental frequency \( f_0 \) and each harmonic \( i \times f_0 \) \((i = 1, 2, 3, \ldots 2^n - 1)\).

2.2. New generator design

Figure 2 indicates that the m-sequence has a broad and configurable passband, making it a suitable signal to simulate the MT source. Accordingly, we designed a broadband signal generator to output the appropriate m-sequence for MT receiver indoor testing. This new generator includes three key modules: a clock module, a PRBS logic module, and a voltage level conversion module, as shown in figure 3. The structure in the dotted box is the clock module which outputs the code rate to
the PRBS logic modules. \(f_b\) is the basic clock frequency and \(f_c\) is the clock rate. PRBS logic modules output two independent m-sequences (unipolar signal) according to \(f_c\) and the register logical structure. The structure in the red box is the voltage level conversion module. This module converts the 3.3 V unipolar signal to an adjustable amplitude bipolar signal. Next, we will detail the design methodology of each module.

2.2.1. Clock module. When designing the clock module, the basic clock should be determined first. Based on the analysis in section 1, we need two m-sequences with a bandwidth covering \(5 \times 10^{-4} - 5 \times 10^2\) Hz and \(0.1 - 1 \times 10^4\) Hz respectively.

An m-sequence is determined by two key indexes: \(n\) and \(f_c\), which permit independent control of the frequency points \(N\) and the bandwidth \(f_{mb}\) (\(n\) is proportional to \(N\) with fixed \(f_c\), and \(f_c\) is proportional to \(f_{mb}\) with fixed \(N\)). When designing the broadband source, we want the frequency range to be as broad as possible to obtain an excellent frequency response. However, these parameters cannot be arbitrarily set, and must be determined in terms of certain rules.

According to equation (3), the m-sequence has maximum power at the fundamental frequency \(f_0\). When \(n \gg 1\), the PSD amplitude at \(f_0\) is:

\[
P_b = A^2 \left( \frac{N+1}{N^2} \right) \approx \frac{A^2}{N} \tag{4}\]

and the \(-3\) dB frequency point \(f_{-3 \, \text{dB}}\) of this m-sequence is calculated by equation (5):

\[
10 \log \left( \frac{P_X}{P_0} \right) = 20 \log \left( \frac{\sin \frac{\pi f_0}{f_c}}{\frac{\pi f_0}{f_c}} \right) = -3 \tag{5}\]

where \(P_X\) is the PSD at \(-3\) dB. Thus,

\[
f_{-3 \, \text{dB}} = k f_0 = 0.4422 \times f_c \tag{6}\]

From equation (6), the \(-3\) dB bandwidth is \(0 - 0.4422 \times f_c\). The PSD in this range is approximate to a straight line, so the m-sequence could serve as white noise for MT receiver testing. Generally, when designing the broadband source, we should ensure that the \(-3\) dB bandwidth is broader than the bandwidth of the testing system \(f_b\). In practice, we should choose \(f_c\) according to the specific application. For example, we need to select a high \(f_c\) for a system with damped oscillation, which needs a very broad testing band, but it is not necessary to select such a large \(f_c\) for a system with low-pass characteristics. The target of the MT method is to measure the electromagnetic structure of the Earth, which is a linear system with low-pass characteristics, and exploration depth increases with decreasing source frequency. The MT receivers have similar properties, hence we selected \(f_c\) for our signal according to equation (7):

\[
0.7 \times f_c \geq f_b \tag{7}\]

Furthermore, the complete MT bandwidth, \(5 \times 10^{-4} - 1 \times 10^4\) Hz, is too broad to realistically test using a single source. Thus, we divided the whole MT bandwidth into two parts to cover the whole MT band: \(5 \times 10^{-4} - 5 \times 10^2\) Hz and \(0.1 - 1 \times 10^4\) Hz, and designed two m-sequences accordingly. Substituting the bandwidths into equation (7) gives:

\[
f_c \geq f_b/0.7 \geq \begin{cases} 500(0.7) = 714.2587\,\text{Hz} \\ 1 \times 10^4(0.7) = 14.286\,\text{kHz} \end{cases} \tag{8}\]

where \(f_b\) is the maximum frequency in the band of interest.

Thus, the scope of \(f_c\) was determined. In the hardware, \(f_c\) is obtained from the basic clock and the frequency divider. Considering equation (8), we choose a 14-stage ripple carry binary counter CD4060 with external RC (resistor–capacitor) oscillation circuits to generate a basic clock rate of 120 kHz, and the clock frequency was obtained from the frequency
divider MAX V 5M80ZE64 CPLD. The frequency dividing depth has two options: 5 and 100. Finally, we made final appropriate values for \( f_c \): 1200 Hz and 24kHz. These parameter choices provide attenuation in the two target bands, both of 2.64 dB, meeting the bandwidth requirement of an MT testing source.

We included a mode exchange button and an indicator LED, so users can choose \( f_c \) for different MT ranges. When the button is depressed, the LED is turned on and the generator is set to NSMT mode with \( f_c = 1200 \) Hz. When the button is released, the LED is turned off and the generator is set to AMT/CSAMT mode with \( f_c = 24 \) kHz. The control circuit delivers the division ratio based on user-selected operations, and the corresponding \( f_c \) is output to the PRBS logic module.

2.2.2. PRBS logic module. The PRBS logic module outputs two binary signals for mode \( xy \) (Ex and Hy) and mode \( yx \) (Ey and Hx). A PRBS is produced by a linear shift register and the logic structure of the register sequence is determined by the primitive polynomial. Before primitive polynomial selection, \( n \) should be determined.

As the sequence length increases, the PRBS spectrum flattens for increasingly higher frequencies as it approaches the purely flat spectrum of an impulse (Pangratz and Weinrichter 1979). To guarantee sufficient frequency resolution, \( n \) should be determined by

\[
\frac{f_c}{2^n - 1} \leq \frac{f_{\Delta}}{(1.25 \sim 1.5)}.
\]  

(9)

where \( f_{\Delta} \) is the minimum frequency in the band of interest.

Increasing \( n \) causes a strong decrease in the source amplitude. However, as we show later, this loss in source energy can be compensated for by a concomitant increase in stacking length, as shown in figure 4, for fixed \( f_c \).

With increasing \( n \), PSD becomes continuous. Thus, larger \( n \) provides a PRBS closer to actual white noise. However, the measurement period also increases with \( n \), which leads to a rapid increase in sample quantity required to guarantee precision. Consequently, the computation to calculate the PSD becomes more complex as well, the signal power allocated to each frequency reduces rapidly (Davidson et al. 2014), and the SNR declines sharply. Thus, to ensure an adequate SNR, \( n \) should not be too large. Considering these issues simultaneously, we defined the selection criteria for \( n \) as equation (10):

\[
(2^n - 1) = (1.25 \sim 1.5) \times (4 - 5)T_s
\]

\[
2^n - 1 = 7.5 \times T_s = 7.5 \times \frac{f_c}{f_{\text{min}}} \Rightarrow n = \left\lfloor \log_2 \left( \frac{1200 \times 7.5}{5 \times 10^{-4}} \right) \right\rfloor = 25
\]  

(10)

where \( f_c \) is the clock rate, and the operation \( \lfloor \rfloor \) means rounding up to the nearest integer.

Once \( n \) is determined, the primitive polynomial of the m-sequence can be obtained from a look-up table (Zivkovic 1994). We selected the primitive polynomial as:

\[
f_{52}(x) = 1 \oplus x^7 \oplus x^{25}, \quad \text{where } \oplus \text{ is modulo-two addition.}
\]

We can obtain different m-sequences using the same generator polynomial with different initial states. The initial state can be anything except all \( \{0\} \), which will cause an endless loop. To achieve strong auto-correlation and weak cross-correlation, we designed two initial states: \( \{1\} \) (channel 1 and 2) and \( \{\text{zeros (1, 24), 1}\} \) (channel 3 and 4), providing a large phase difference between the two m-sequences. The schematic of the PRBS logic module is shown in figure 5.

2.2.3. Voltage level conversion module. The sequence output by the PRBS logic module is CMOS level, with amplitude 3.3V, including the DC component. However, the broadband source must be a bipolar signal. The voltage level conversion module provides this transformation by a means of a series, including comparator, amplifier and attenuator. The binary sequence was used as the control signal; \( V_{\text{Ref}} \) with precision 0.01%, as a reference signal; and a unipolar analog signal was produced by an analog switch with voltage equal to \( V_{\text{Ref}} \) and phase equal to the binary PRBS. The bipolar signal was obtained by subtracting \( V_{\text{Ref}} \) from the unipolar signal. Finally, the standard signal source was generated through the attenuator circuit (box ‘Att’ within the structure in the red box in figure 3).
Overall, the generator produces four signals: channels 1 and 2 are the same signals with different amplitude, as are channels 3 and 4. The relationships between the output channels and MT signal source are:

- channel 1 = \(E_x\) (±1 mV bipolar PRBS),
- channel 2 = \(H_y\) (±5 mV bipolar PRBS),
- channel 3 = \(E_y\) (±1 mV bipolar PRBS),
- channel 4 = \(H_x\) (±5 mV bipolar PRBS).

Channel 1 and 2 are used to test MT mode \(xy\), and channel 3 and 4 are used to test MT mode \(yx\).

A photograph of the developed signal generator is shown in figure 6.

Theoretically, the wave forms of the signals obtained by this generator are shown in figures 7 and 8. Figure 7 shows the source wave in the time domain. Channel 1 (\(E_x\)) and channel 2 (\(H_y\)) (mode \(xy\)) have the same wave shape with different amplitude \(A\), and similarly for channel 3 (\(E_y\)) and channel 4 (\(H_x\)) (mode \(yx\)). The two PRBSs of different \(f_c\) should be in accordance with each other, and the only difference is the value of \(\Delta t\). Figure 8 shows the PSD for the PRBS. The PSD is similar to white noise, and the effective frequencies denoted by the cycle points scatter around a straight line in the corresponding band.

3. Tests

We conducted tests to verify the performance of the new generator including effective bandwidth and application effect.
The tests were divided into two parts: (1) indoor testing: to ascertain the frequency response of a new MT receiver developed by China University of Geosciences (Beijing) (CUGB) by using the source output by the new generator in-lab and obtain the calibration coefficients; (2): field testing: to verify the working ability of the new receiver which was tested using the new generator.

For indoor testing, we used the new generator as the signal source and an MT receiver (newly developed by CUGB) as the data logger. Switching $f_c$, we performed data acquisition and tested for NSMT and AMT/CSAMT ranges. Sampling times $T$ were six hours, two hours and ten minutes with sampling rates 150 Hz and 2400 Hz and 24 kHz respectively. The results are shown in figures 9–12.

We only plotted the signals with $f_c = 1200$ Hz and $f_c = 24$ kHz for clarity. The wave forms with other parameters were a similar shape, but with a different pulse width.

The auto-correlation was calculated using $Ex$ and the cross-correlation of different channels. The auto-correlation of $Ex$ and cross-correlation of different channels.

\[
H(\omega) = \frac{R_1 [R_2(1 + j\omega(C_1 + C_2)R_3) + R_3]}{(R_1 + R_2)(1 + j\omega(C_1 + C_2)R_3) + R_3} \\
= \frac{R_2(R_2 + R_3)(R_1 + R_2 + R_3)}{(R_1 + R_2 + R_3)^2 + (C_1 + C_2)^2(R_1 + R_3)^2} \\
+ j\frac{\omega(C_1 + C_2)}{(R_1 + R_2 + R_3)^2 + (C_1 + C_2)^2(R_1 + R_3)^2}. \\
\]

(11)

The 25-order PRBS described above with register logic of $\{1\}$ was used as an input stimulus current. Three sample frequencies were used to collect the raw voltage data: 15 Hz, 150 Hz, and 2400 Hz. Then, we carried out a fast Fourier transform for the excitation and response data. Finally, the frequency response of the model was obtained.

A comparison of the calculated frequency response and the theoretical values is shown in figure 14. These were joined by three sections of different $f_c$.

Figure 14 indicated that the frequency response of a system could be described accurately by using the newly designed broadband signal.

We used the new generator to test and calibrate the new MT receiver. We selected a large current to guarantee enough SNR, so the reference noise was omitted from consideration. The testing results are shown in figure 15.

Indoor testing indicated that, the new designed signal generator could test the MT receiver across the whole MT band. Results show that the MT receiver has a stable response in the target bands. After indoor testing, field tests were conducted to evaluate the application performance of the new generator. The test site was in Zhangbei County in the north of China and comparison tests were conducted between the tested receiver and the MTU 5 receiver from Phoenix. Two days’ worth of MT data was obtained.

Phoenix is a geophysical instrument company that produces apparatus which is among the best in the world. Therefore, we can assume that the results obtained by MTU 5 are accurate and can be used as the measuring criteria. The apparent resistivity and impedance phase results are shown in figure 16. The blue line and the red line represent the results before and after revision by the calibration coefficients, respectively. They are obtained by the newly
designed receiver from CUGB, and the black line represents the results from MTU 5. During the data processing, frequency-domain interpolation was used to calculate the correction coefficients at the corresponding target frequencies. Then, we processed the amplitude and phase by dividing the PSD of obtained signals. The relationships between the PSD and line types are: black lines are PSD of the mode $xy$; red lines are PSD of mode $yx$; solid lines are PSD of the electric field; dotted lines are PSD of the magnetic field.

Figure 12. Coherence of the four-channel signals. $Coh_{ExHy}$ (blue line) represents the coherence of $Ex$ and $Hy$ (channel 1 and 2); $Coh_{EyHx}$ (black line) represents the coherence of $Ey$ and $Hx$ (channel 3 and 4); $Coh_{ExEy}$ (red line) represents the coherence of $Ey$ and $Ey$ (channel 1 and 3); $Coh_{HxHy}$ (green line) represents the coherence of $Hy$ and $Hx$ (channel 2 and 4).

Figure 13. Simulink model for calibration reference.

Figure 14. Frequency response of the simulation model. Symbols: ‘O’ is the theoretical values and ‘*’ is the calculated values.
and subtracting the calibration coefficients, respectively. Thus, the revised results were obtained. Obviously, the receiver could carry out field MT measurement successfully, which is consistent with the indoor testing using the new generator. Compared with the unrevised results, the revised results dealt with the calibration coefficients obtained in-lab by using the broadband signal generator were closer to the MTU 5 results.

4. Conclusion

In the development of a new MT receiver, indoor and field tests must be conducted to evaluate the performance and guide improvements. Because of the large bandwidth of MT methods, white noise testing is a useful tool to establish the precision and frequency-response character of the MT receiver. However, currently there is no suitable source generator to perform such testing. Accordingly, we have designed a broadband signal generator based on PRBSs to simulate an MT field in a variety of bandwidths. Desirable bandwidths are different for different MT methods, so we divided the whole band into two segments: $5 \times 10^{-4} - 5 \times 10^2$ Hz and $0.1 - 1 \times 10^4$ Hz, and determined the key indexes and the primitive polynomial. The new signal generator consists of three parts: a clock module, a PRBS logic module, and a voltage level conversion modules. This generator can provide two-mode (mode $xy$: channel 1 and 2, mode $yx$: channel 3 and 4) independent bipolar broadband signals depending on the parameters selected by users.

Indoor tests show that the new signal generator provides independent broadband signals with the spectral characteristic of white noise. The PSD of the broadband source is very flat across a broad frequency band, and the frequency responses of the four channels are consistent with each other. Field tests proved the necessity of testing and calibrating a new MT receiver in-lab.

The new generator is very useful for MT receiver testing, and it may also be used for other frequency domain methods. However, there remain some issues with low frequency due to sensitivity loss in the sensor. Thus, the amplitude of the induced voltage decreases, and large errors arise from interference. We can improve this situation in the future by increasing the SNR and further appropriate data processing.

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