



## 1 **Multifunction waveform generator for EM receiver** 2 **testing**

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6 **Abstract.** In many electromagnetic (EM) methods, such as magnetotelluric, spectrum induced  
7 polarization, time domain induced polarization, and controlled source audio magnetotelluric methods,  
8 it is important to evaluate and test the EM receivers during their development stage. To assess the  
9 performance of the developed EM receivers, controlled synthetic data that simulates the observed  
10 signals in different modes is required. Based on our testing, the frequency range, frequency precision,  
11 and time synchronization of the currently available function waveform generators in the market are  
12 deficient. This paper presents a multifunction waveform generator with three waveforms: 1) a  
13 wide-band low-noise electromagnetic field signal to be used for magnetotelluric, audio-magnetotelluric,  
14 and long period magnetotelluric studies; 2) a repeating frequency sweep square waveform for  
15 controlled source audio magnetotelluric and spectrum induced polarization studies; and 3) a  
16 “positive-zero-negative-zero” signal that contains primary and secondary fields for time domain  
17 induced polarization studies. In this paper, we provide the principles of the above three waveforms  
18 along with a hardware design for the generator. Furthermore, testing of the EM receiver was conducted  
19 with the waveform generator, and the results of the experiment were compared with those calculated  
20 from the simulation and theory in the frequency band of interest.

21  
22 **Keywords:** Multifunction waveform generator; EM receiver; pseudo-random binary sequence; chopper;  
23 signal synthesiser

### 24 **1 Introduction**

25 Electromagnetic (EM) methods are successfully used in a variety of applications, including metal ore  
26 investigations, ground water explorations, hydrocarbon prospecting, volcano research, and deep earth  
27 research (Wei et al., 2010; Key, 2003). In these applications, EM methods are used to measure natural



28 or controlled source signals, and then to interpret the underground electrical structure using data  
29 processing and inversion techniques (Osinowo and Olayinka, 2012; Scheuermann, 2016). There are  
30 many EM methods in use today; however, in this paper we are interested in magnetotelluric (MT)  
31 (Cagniard, 1953), controlled source audio magnetotelluric (CSAMT) (Sandberg and Hohmann, 1982),  
32 spectrum induced polarization (SIP) (Johnson, 1984), and time domain induced polarization (TDIP)  
33 (Marshall and Madden, 1959) methods. The requirements for field instruments supporting these  
34 methods include high resolution, large exploration depth, low cost, and high efficiency field data  
35 acquisition. All of the above EM methods are dependent on the quality of the field data acquired by the  
36 EM instrument. Therefore, the specifications of the EM receiver are of particular importance for high  
37 quality EM prospecting. The current EM receivers, such as the V8 receiver from Phoenix Geophysics  
38 (PhoenixGeophysics, 2017), GDP32 from Zonge (Zonge, 2017), ADU-07e from Metronix (Metronix,  
39 2017), and KMS-820 from KMS (KMS, 2017), are all specified as being multifunction, multi-channel,  
40 and easy to use with low noise levels and low clock drift errors.

41 The China University of Geoscience began developing a multifunction EM receiver (EMR6) in 2014,  
42 which is mainly intended for EM field measurements in surface and tunnel. The receiver support  
43 method modules include the MT, CSAMT, SIP, and TDIP methods. In addition, the MT function  
44 includes the audio magnetotelluric (AMT), magnetotelluric, and long period magnetotelluric (LMT)  
45 methods in different frequency ranges. During the development of the instrument, the receiver was  
46 tested with each method module after the electrical parameters were characterised. For the MT method  
47 module test (Ge et al., 2016), a pseudorandom bit sequence (PBRs) module was designed, which  
48 included a white noise source that simulated a broadband natural source MT signal in the  $5 \times 10^{-4}$  Hz to  
49 14 kHz frequency range, which was suitable for the AMT and MT mode. However, the lower  
50 frequency range for the LMT test was deficient. For the CSAMT and SIP mode tests, which require a  
51 repeating frequency sweep square waveform, the test signal should be time synchronized and repeated.  
52 There are function waveform generators available in the market, such as the Agilent 33250A, which  
53 support multiple broadband waveforms; however, the frequency precision and time synchronization  
54 error in these generators are deficient. For example, in the TDIP mode test, the arbitrary waveform  
55 function is suitable, but the time synchronization error is deficient.

56 We developed a special multifunction waveform generator to meet the testing requirements of the  
57 EMR6 receiver. The waveform generator supports a broadband low-noise pseudo-random binary



58 sequence (PRBS) for MT, repeating frequency sweep square waveform for CSAMT and SIP with a  
59 programmable frequency step list, and “positive-zero-negative-zero” (PNZN) waveform containing  
60 both primary and secondary fields for the TDIP mode. Moreover, additional requirements for the  
61 waveform generator included: 1) ease of use, auto repeat circulate output according the scheduled  
62 frequency step list; 2) high precision phase and time synchronisation; 3) low power consumption for  
63 the LMT mode test, which has the capacity of the built in Li-ion battery that allows for one week of  
64 operation.

65 This paper presents the details of the generator, including both the system design and the hardware  
66 principles. The results of the experimental testing using the EMR6 are also provided.

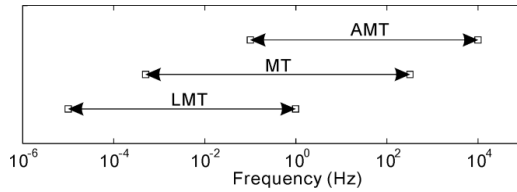
## 67 **2 System design**

68 The multifunction waveform generator is designed to output three types of waveforms: white noise for  
69 MT (AMT/MT/LMT), a repeated swept square waveform for CSAMT and SIP modes, and a “PNZN”  
70 waveform for the TDIP mode.

71 The MT method consists of three branches – AMT, standardised MT, and LMT – and these branches  
72 differ in terms of the exploration depths of interest and the effective frequency ranges. Figure 1  
73 illustrates the frequency ranges applicable to the three branches of the MT method. The high frequency  
74 band in the AMT method is from 10 kHz to 0.1 Hz, which spans five decades; the frequency band in  
75 the standardised MT method is from 320 Hz to  $5 \times 10^{-4}$  Hz, which spans approximately six decades;  
76 the frequency band in the LMT method is from 1 Hz to  $1 \times 10^{-5}$  Hz, which spans approximately five  
77 decades. A PRBS generator is the best choice for generating broadband signals (Amrani et al., 1998).

78 The highest frequency is determined by the width of the smallest encoding, and the lowest frequency  
79 depends on the length of the PRBS. According to the three different frequency band modes, the length  
80 of the PRBS must be greater than  $1 \times 10^6$ . Therefore, we designed a PRBS with length  $2^N - 1$ , where N  
81 equals 24 and the length is 16 M. By changing the smallest code width of the PRBS, it was easy to  
82 meet the three MT modes operating in different frequency ranges.

83



84  
 85 **Figure 1: Three MT sounding methods with different frequency ranges.**

86 The formal definition of the PRBS is:

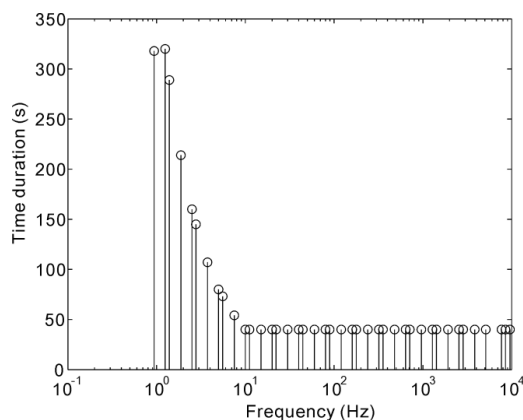
87 
$$D = X^{24} + X^7 + X^2 + X + 1. \quad (1)$$

88 The length of the sequence is 16 M. The smallest code width, which changes for different fundamental  
 89 frequencies, can be selected from 10  $\mu$ s, 1 ms, and 100 ms. Table 1 lists the parameters for the three  
 90 modes.

91  
 92 **Table 1. PRBS Parameters for three MT modes.**

Mode	Length	Smallest code width	Available frequency range
AMT	16 M	10 $\mu$ s	100 kHz to 6 mHz
MT	16 M	1 ms	1 kHz to 60 $\mu$ Hz
LMT	16 M	100 ms	10 Hz to 0.6 $\mu$ Hz

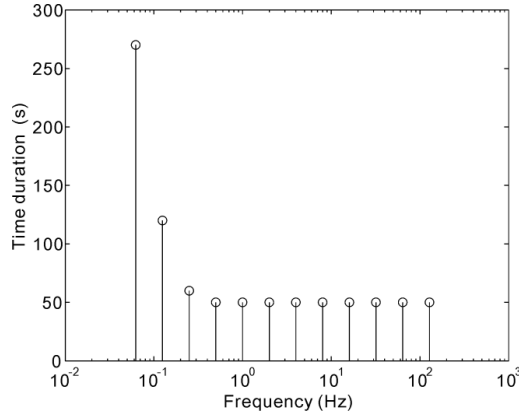
93  
 94 The designed frequency range for the CSAMT method is from approximately 0.1 Hz to 10 kHz. The  
 95 programmable frequency stepping schedule is according to the typical CSAMT frequency stepping  
 96 schedule used in the field when working with V8 receiver from Phoenix Geophysics. The frequency  
 97 stepping schedule contains 41 frequency points, and lasts for 50 min. Figure 2 presents the typical  
 98 frequency stepping schedule plotted on a log scale, and two frequency points in a double frequency  
 99 range as an approximate average in the log scale. The 41 frequency points cover four decades and  
 100 extend from 9,600 Hz to 0.9375 Hz. All frequency points are derived from a 12.288 MHz high-stability  
 101 clock source. To increase the lower frequency signal-to-noise ratio (SNR), the length of the stacking  
 102 time of the lower frequencies is longer than that of the higher frequencies. The longest stacking time is  
 103 323 s for 1.25 Hz in the low frequency band, and the shortest is 40 s in the high frequency band.  
 104



105

106 **Figure 2: Stacking time length for a typical CSAMT frequency stepping schedule.**

107 The designed frequency range for the SIP method is from 128 Hz to 0.0625 Hz. The frequency stepping  
108 schedule is the same as the typical SIP step schedule used in the field when working with V8 receiver  
109 from Phoenix Geophysics. Figure 3 shows a typical frequency stepping schedule plotted on a log scale,  
110 and frequency stepping by double. There are 12 frequency points, each of which last for 15 min, and  
111 the 12 points cover approximately four decades and extend from 128 Hz to 0.0625 Hz. As was the case  
112 in the CSAMT mode, all frequency points could be derived from a 12.288 MHz clock source. To  
113 increase the lower frequency SNR, a longer stack time was used in the lower frequency band versus  
114 that used in the higher frequency band. The longest stack time was 273 s in the low frequency band at  
115 1.25 Hz, and each point lasted for 50 s in the high frequency band.



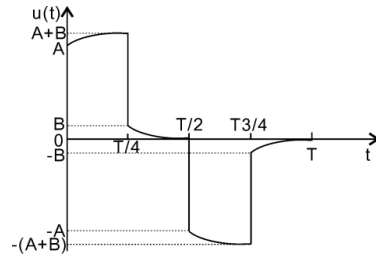
116

117 **Figure 3: Stack time length for a typical SIP frequency stepping schedule.**

118

119 The PZNZ waveform for the TDIP module is shown in Fig. 4 and consists of four phases: positive ON  
 120 time, OFF time, negative ON time, and OFF time. The duty ratio is 1:1. In the figure, T denotes the  
 121 four-phase period and the width of the pulse is T/4. The term A denotes the amplitude of the primary  
 122 electrical field, B denotes the maximum amplitude of the secondary electrical field, and A + B denotes  
 123 the total electrical field. The self-potential and other disturbances are not considered in the figure.

124



125

126 **Figure 4: PZNZ waveform from a theoretical simulation.**

127

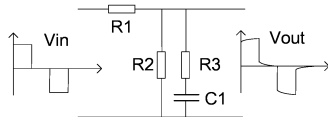
$$128 \quad u_1(t) = \begin{cases} A, & 0 \leq t < T/4 \\ 0, & T/4 \leq t < T/2 \\ -A, & T/2 \leq t < 3T/4 \\ 0, & 3T/4 \leq t < T \end{cases} \quad (2)$$

$$129 \quad u_2(t) = \begin{cases} B - Be^{-t/\tau}, & 0 \leq t < T/4 \\ Be^{-(t-T/4)/\tau}, & T/4 \leq t < T/2 \\ Be^{-(t-T/2)/\tau} - B, & T/2 \leq t < 3T/4 \\ -Be^{-(t-3T/4)/\tau}, & 3T/4 \leq t < T \end{cases} \quad (3)$$



130 where  $u_1(t)$  and  $u_2(t)$  denote the primary and secondary electrical fields, respectively, and the exponent  
 131 attenuation curve refers to the secondary electrical field.

132  
 133



134

135 **Figure 5: Schematic of the PZNZ waveform generator.**

136 Figure 5 shows a schematic of the circuit used to generate the PZNZ waveform, which contains both  
 137 primary and secondary electrical fields. The resistor-capacitor (RC) network transforms the PZNZ ( $V_{in}$ )  
 138 waveform into a PZNZ waveform ( $V_{out}$ ) with the secondary electrical field.

139

140 
$$A = \frac{R_3 // R_2}{R_1 + R_3 // R_2} K \quad (4)$$

141 
$$B = \frac{R_3 // R_2}{R_1 + R_3 // R_2} K - \frac{R_2}{R_1 + R_2} K \quad (5)$$

142 
$$\tau = R_3 C_1 \quad (6)$$

143

144 The amplitude of  $V_{in}$  is  $\pm K$ . A and B used in Eqs. (2) and (3) can be calculated from Eqs. (4) and (5).

145 The time constant ( $\tau$ ) is given by Eq. (6).

146 **3 Hardware principle**

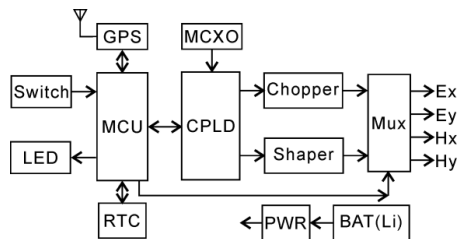
147 **3.1 Block diagram**

148 Figure 6 presents a block diagram of the waveform generator, which has the advantages of automation,  
 149 ease of use, high phase precision, and low power. As shown in the figure, the hardware consists of an  
 150 input switch, microcontroller unit (MCU), GPS module, complex programmable logic device (CPLD),  
 151 compensated microprocessor crystal (MCXO), real time clock (RTC), chopper, shaper, multiplexer,  
 152 power conversion circuit, and built-in Li-ion battery package. The input switch is used to change the  
 153 work mode without configuring the external complicated parameters. The LED is used to indicate the  
 154 working status. The different work modes output different LED flash patterns. The GPS module is  
 155 LEA-6T from U-blox, which provides a high precision time pulse per second with low power



156 consumption. The MCXO has the specifications of high stability clock source (12.288 MHz) ( $\pm 30$  ppb)  
157 and low power consumption (3.3 V & 12 mA). To lock the GPS, the MCU receives the time  
158 information from the GPS module and writes to the RTC. The CPLD is used to implement a frequency  
159 divider, logic operator, PPS lock, and tracking. The RTC is the time counter used for circulation of the  
160 frequency stepping schedule. The chopper circuit chops a high precision DC reference into a bipolar  
161 square waveform under the control of the CPLD. The shaper generates the TDIP PZMZ waveform.  
162 Moreover, the sum of the primary and secondary field signals is also provided as output. The  
163 multiplexer is controlled by the MCU to select either the chopper or the shaper output. The power  
164 module converts the Li-ion battery (11.1 V and 10 Ahr) voltage to digital power at 3.3 V and analogue  
165 power at  $\pm 3$  V.

166



167

168 **Figure 6: Block diagram of the multifunction waveform generator circuit.**

169 When the MT mode is selected, the MCU controls the CPLD to generate two independent PRBS. The  
170 outputs Ex and Hy share one PRBS, whereas Ey and Hx share another PRBS. The two PRBSs are  
171 different for different phases. The chopper circuit converts the 3.3 V COMS PRBS into a bipolar  $\pm 10$   
172 mV square wave for Ex and Ey. The amplitude of Hx and Hy is  $\pm 100$  mV. The MCU uses the  
173 multiplexer output to select the chopper output. Based on the different modes (AMY/MT/LMT), the  
174 output selected by the input switch and the smallest code width of the PRBS is changed. To decrease  
175 the power consumption, the GPS module is powered down during the MT mode.

176 When the CSAMT mode is selected, the MCU reads the GPS time information and writes to the  
177 RTC. The MCU controls the CPLD division factor according to the frequency stepping schedule.  
178 The CPLD divides the clock source (12.288 MHz) to the target frequency step by step. The divider is  
179 triggered by the PPS from the GPS module. The 41 frequency point clock signal drives the chopper  
180 circuit to generate a bipolar square waveform. All frequency stepping schedules start from the reference  
181 time base of 00:00:00 and repeat from high frequency to low frequency. For example, after the power is





182 turned on and the GPS is locked, the current time is 02:20:00, the period of frequency stepping  
183 schedule is 50 min, the residual time is 40 min, the first output signal is the No. 40 frequency point  
184 (1.25 Hz), and the output is the rest 10 min of the current schedule. The frequency stepping schedule is  
185 repeated continually.

186 The procedure in the SIP mode is the same as that described above for the CSAMT module for the step  
187 schedule with time synchronization. A typical frequency stepping schedule period is 15 min, and  
188 contains 12 frequency points.

189 When the TDIP mode is selected, the RTC time is locked to the GPS module and the CPLD generates  
190 pulse with a 2 s width synchronization as the PPS from the GPS module to drive the shaper circuit. The  
191 reference time base is 00:00:00, and the typical period is 8 s with duty ratio 1:1.

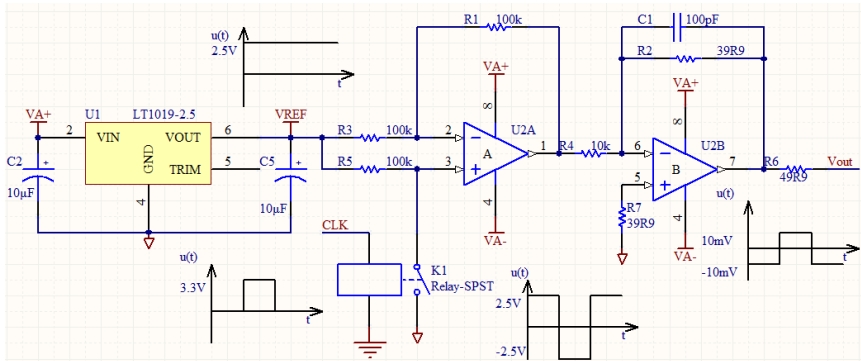
### 192 **3.2 Clock source**

193 In accordance with the high precision phase requirement from the CSAMT and SIP modes, an  
194 integrated U-blox GPS module and MCXO were selected as the high stability clock source. The  
195 LEA-6T module series is a family of stand-alone GPS receivers that feature the high performance  
196 U-blox 6 timing engine. The accuracy of the time pulse signal in the LEA-6T is approximately 30 ns,  
197 and the time-to-first-fix is 29 s. The clock module is an ultra-high stability MCXO from Vectron  
198 MX-503 with an accuracy of  $\pm 30$  ppb in a temperature range of  $-20$ – $75$  °C. The power consumption of  
199 the module is 12 mA at 3.3 V. For example, to generate the clock ( $f = 9,600$  Hz) in the CSAMT mode,  
200 the MCU sets the division factor to 1280 and the CPLD divides the clock signal which is triggered by  
201 the PPS from the GPS module. After 50 PPS counts, the MCU sets the division factor to 1,600, and the  
202 output clock frequency changes to 7,680 Hz. The above steps are then repeated, and each division is  
203 triggered by the PPS.

204 The poorest condition for meeting the high phase precision requirement in the SIP mode is that the  
205 error is 30 ppb during the 15 min circulation, which will cause a time drift of 27  $\mu$ s. This time drift  
206 results in a 21 mrad phase error at 128 Hz.



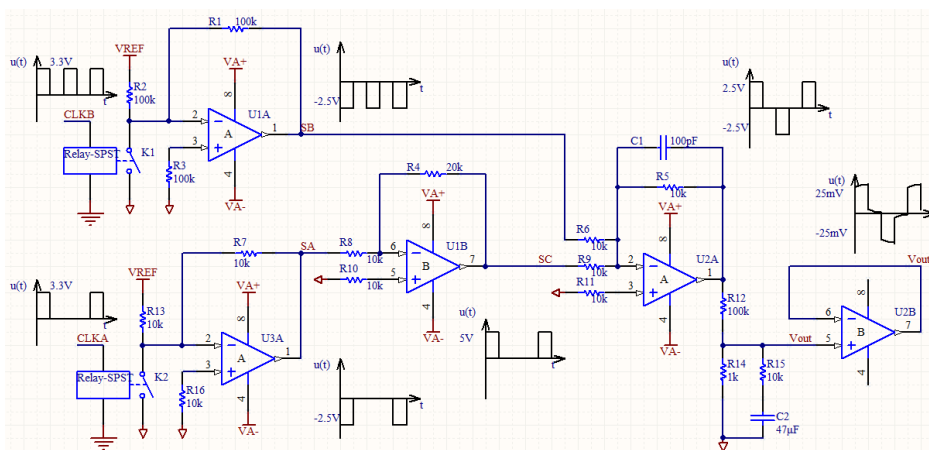
207 **3.3 Chopper circuit**



208  
 209 **Figure 7: Schematic of the chopper circuit (E channel).**

210 Figure 7 shows the schematic of the chopper circuit, which chops a high precision DC reference  
 211 voltage into a bipolar square waveform driven by the switching clock. The circuit contains voltage  
 212 references, a relay-SPST, and an amplifier. The voltage reference LT1019-2.5 (from Linear Technology)  
 213 used is a 2.5 V high precision DC reference. The relay is driven by the switching clock, which is a 3.3  
 214 V CMOS square waveform. The switching clock control the connection U2 Pin3 and GND. The output  
 215 from U2 Pin 1 is a bipolar square wave with an amplitude of 2.5 V. The component U2B is an  
 216 attenuator and low pass filter, and the output decreases from 5 Vpp to 20 mVpp for the E channel. The  
 217 bandwidth is limited to 100 kHz. The amplitude of the H channel is 200 mVpp, which is different from  
 218 the gain of U2B.

219 **3.4 Shaper circuit**



220



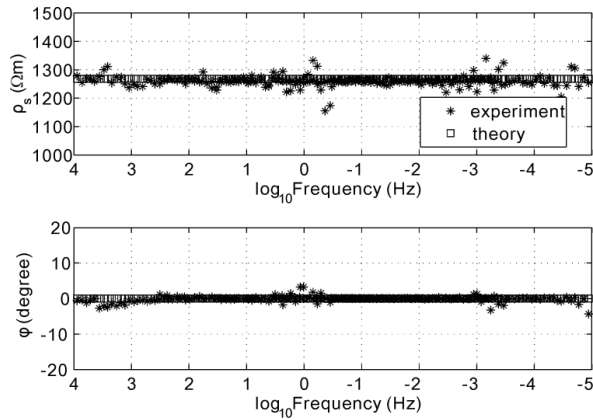
221 **Figure 8: Schematic of the shaper circuit.**

222 A schematic of the shaper circuit is shown in Fig. 8. The shaper circuit shapes the clock output from the  
223 CPLD into a PZMZ waveform. In the shaper circuit, there are two channel clock sources (CLKA and  
224 CLKB) and one output (Vout). The CLKA and CLKB outputs are set as different frequencies such that  
225 the frequency of CLKA is double the frequency of CLKB, and the duty ratio of CLKB is 1:3. The  
226 width of each pulse is 2 s. The amplifiers U1A and U3A are both used as those in the previous chopper  
227 circuit, which chops the DC reference to the square waveform. The clock is transformed into a square  
228 waveform with an amplitude of 2.5 V. The amplifier U1B is set with a gain of  $-2$ . The amplifier U2A  
229 adds nodes SB and SC. The waveform at node SD is a bipolar PZMZ waveform. The components R12,  
230 R14, R15, and C2 constitute an RC network to generate the PZMZ waveform based on the principles  
231 described earlier. The amplitude at the output of the shaper is approximately 50 mVpp.

232 **4 Test**

233 **4.1 MT mode**

234 We used the developed multifunction waveform generator to test our multifunction EM receiver  
235 (EMR6). The output of the multifunction waveform generator was connected to the input of the EMR6,  
236 which works in three sequential modes: 30 min for AMT mode, 24 h for MT mode, and 72 h for LMT  
237 mode. All raw data was processed using the white noise method by the MT data processing software  
238 (SSMT2000) from Phoenix Geophysics. The results of the data processing are shown in Fig. 9. The  
239 apparent resistivity and impedance phase are present across the entire nine-decade frequency range  
240 from  $1 \times 10^{-5}$  Hz to  $1 \times 10^4$  Hz. The theoretical simulation result of the apparent resistivity was 1,270  
241  $\Omega$  m, and the impedance phase was approximately  $0^\circ$ . When the experimental results were compared to  
242 those from the theoretical simulation, the bias error between the experiment results and theoretical  
243 value for the apparent resistivity was 1.5%.  
244



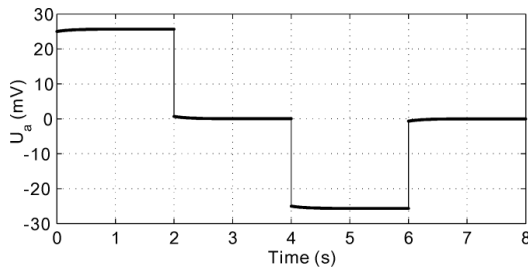
245

246 **Figure 9: Apparent resistivity and impedance phase results from the EMR6 testing. The upper subfigure**  
247 **shows the apparent resistivity curves, and the lower subfigure shows the impedance phase curves.**

248 The results of the experiment indicated that the multifunction waveform generator could test the MT  
249 receiver across the entire MT band, and the EMR6 receiver was shown to have a correct response in the  
250 target bands.

#### 251 4.2 TDIP mode

252 To verify the TDIP waveform, the generator was switched to the TDIP mode and its output was  
253 connected to the EMR6 E channel input, and the EMR performs TDIP data acquisition during 2 min at  
254 2,400 Hz sample rate. Figure 10 shows the time series captured by the EMR6. The full waveform was  
255 recorded for the entire time series of the E1 channel, including the primary and secondary field  
256 waveforms. The amplitude was approximately 50 mVpp, and the “on” time and “off” time pulse widths  
257 were 2 s.



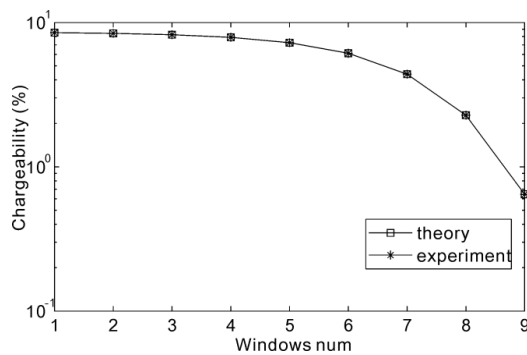
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259 **Figure 10: Time series in TDIP mode as captured by the EMR6.**

260 The chargeability was calculated by dividing the secondary field into nine windows. The offset time of  
261 the first window was 10 ms, and each window had a width of 8, 16, 32, 64, 128, 256, 512, 1024, and



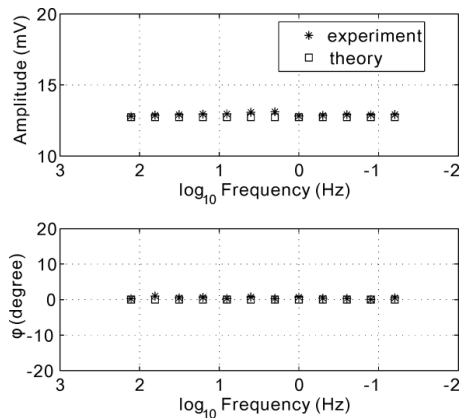
262 2048 sample points, respectively. Figure 11 shows a comparison between the measured results and  
 263 those from the theoretical simulation. The chargeability was distributed from 0.7% to 9%. The results  
 264 show that the output from the waveform generator during the experiment response was in good  
 265 agreement with the theoretical output. The bias error of the nine windows was 0.8% between the  
 266 measured results from the experiment and those from the theoretical simulation.  
 267



268  
 269 **Figure 11: Comparison between the calculated results and the theoretical simulation results for TDIP mode.**

270 **4.3 SIP mode**

271 The EMR6 and waveform generator operated on the SIP mode. To simplify the experiment, the current  
 272 data of the transmitter and a variety of geometric factors were not considered. The EMR6 recorded the  
 273 frequency-swept square waveform for 15 min and calculated the amplitude of each target frequency.  
 274 Figure 12 shows the 12 target frequencies from 128 Hz to 0.0625 Hz. The amplitudes measured in the  
 275 experiment were approximately 12.74 mV and the phases were approximately 0°. The bias error  
 276 between the experiment and theory was 0.5 % across the entire frequency range.



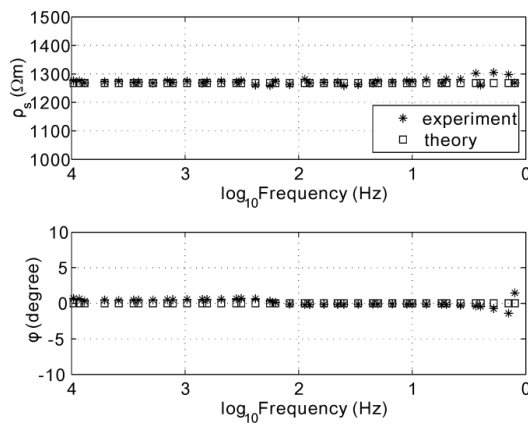


278 **Figure 12: Comparison between the experimental results and the theoretical simulation results for the SIP**  
 279 **mode.**

280 **4.4 CSAMT mode**

281 The EMR6 and waveform generator operated on the CSAMT mode and recorded the E- and H-channel  
 282 swept square waveforms. The Cagniard apparent resistivity and impedance phase were calculated. The  
 283 receiver recorded the swept frequency waveform for 50 min. Figure 13 shows the results calculated for  
 284 the 41 target frequency points from 9,600 Hz to 0.9375 Hz. The apparent resistivity measured in the  
 285 experiment was approximately 1,268  $\Omega$  m and the phase was approximately 0°. The bias error between  
 286 the experiment and theory was 1.3% across the entire frequency range.

287



288

289 **Figure 13: Comparison between the calculated results and the simulation results for the CSAMT mode.**

290 **4.5 Comparison with commercial function waveform generator**

291 From the results of the above experiment, Table 2 presents the comparison of the specifications of the  
 292 developed multifunction waveform generator and Agilent 33510B function waveform generator. The  
 293 presented results indicate that Agilent 33510B is not suitable for EM receiver testing. The developed  
 294 multifunction waveform generator is a better signal source for this purpose.

295

296 **Table 2. Specification comparison with Agilent 33510B.**

Specifications	Agilent 33510B	Newly developed multifunction waveform generator
Function	Sine/square/ramp/pulse/triangle/PRBS/white noise	PRBS/Frequency sweep/PZNZ
Channels	2	4 (Ex/Ey/Hx/Hy)



Time sync	Internal timer or Ext Trig connector	GPS
PRBS	1m bps~50 M bps	10 $\mu$ Hz~100 kHz
Power	AC 100~240 V	Built-in rechargeable Li-ion battery

## 297 **5. Conclusion**

298 The performance of the existing commercial function/arbitrary waveform generator is deficient with  
299 regard to time synchronization and waveform requirements. The multifunction waveform generator  
300 described in this paper was found to be useful for conducting EM receiver testing for multiple EM  
301 methods. The results of the testing show that the multifunction waveform generator could provide three  
302 mode signals containing independent broadband signals with different spectral characteristics, white  
303 noise, a repeating swept square waveform, and a PZMZ waveform consisting of primary and secondary  
304 fields. The apparent resistivity and impedance phase of the broadband white noise source was very flat  
305 across a wide frequency band. The theoretical design of the TDIP waveform was shown to have a  
306 correct response. In the SIP and CSAMT modes, the results of the experiment and theoretical  
307 simulation were a close match. In addition, the generator had other advantages in that it was easy to use  
308 and had low power consumption.

309 Furthermore, various parameters, such as the programmable frequency step schedule in the CSAMT  
310 and SIP modes and the pulse width of the TDIP waveform, are configurable by the user through the  
311 hardware interface or by loading the memory.

## 312 **Author Contributions**

313 Kai Chen developed the required hardware and software. Sheng Jin created the overall design and  
314 performed the tests. Ming Deng was the chopper and shaper circuit technology consultant.

## 315 **Competing Interests**

316 The authors declare that they have no conflict of interest.

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319 of China (2014AA06A603), National Science Foundation of China (61531001), Central University



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